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The Proceedings

OF

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PART B

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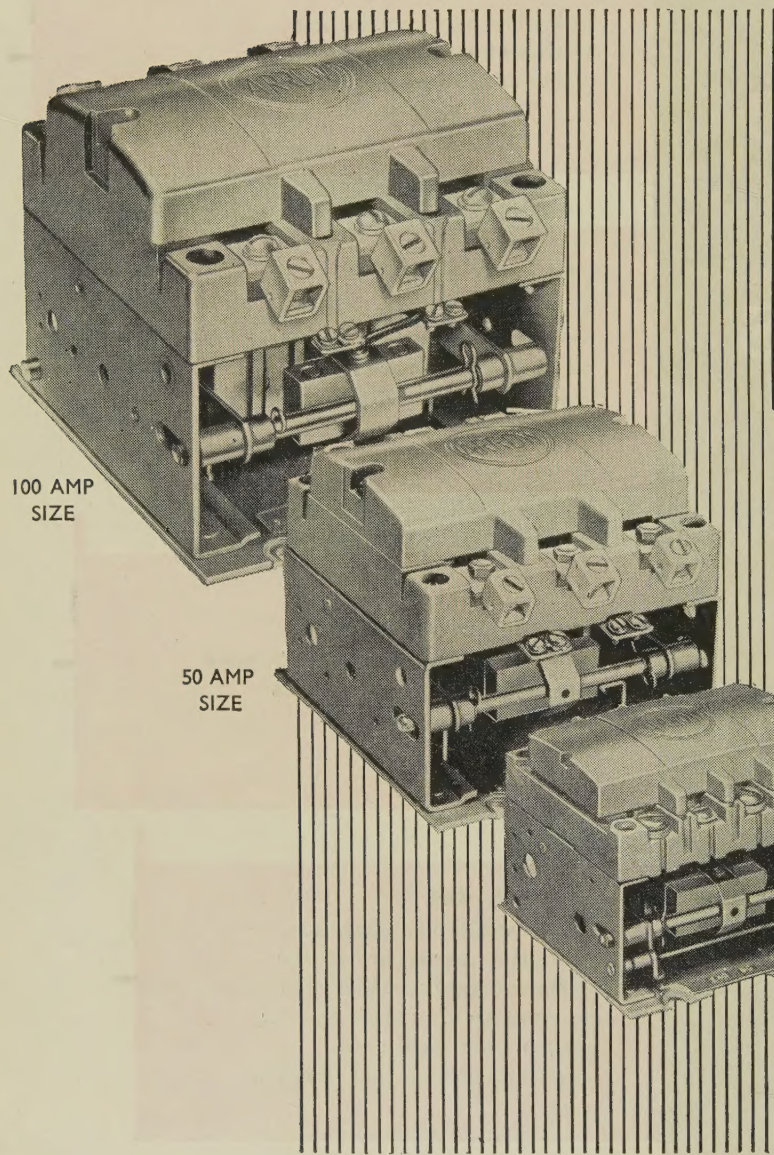
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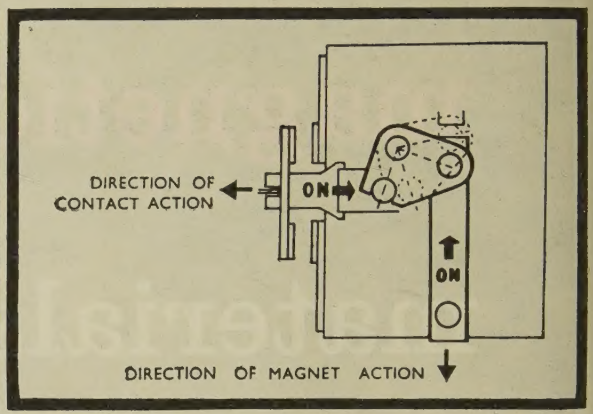
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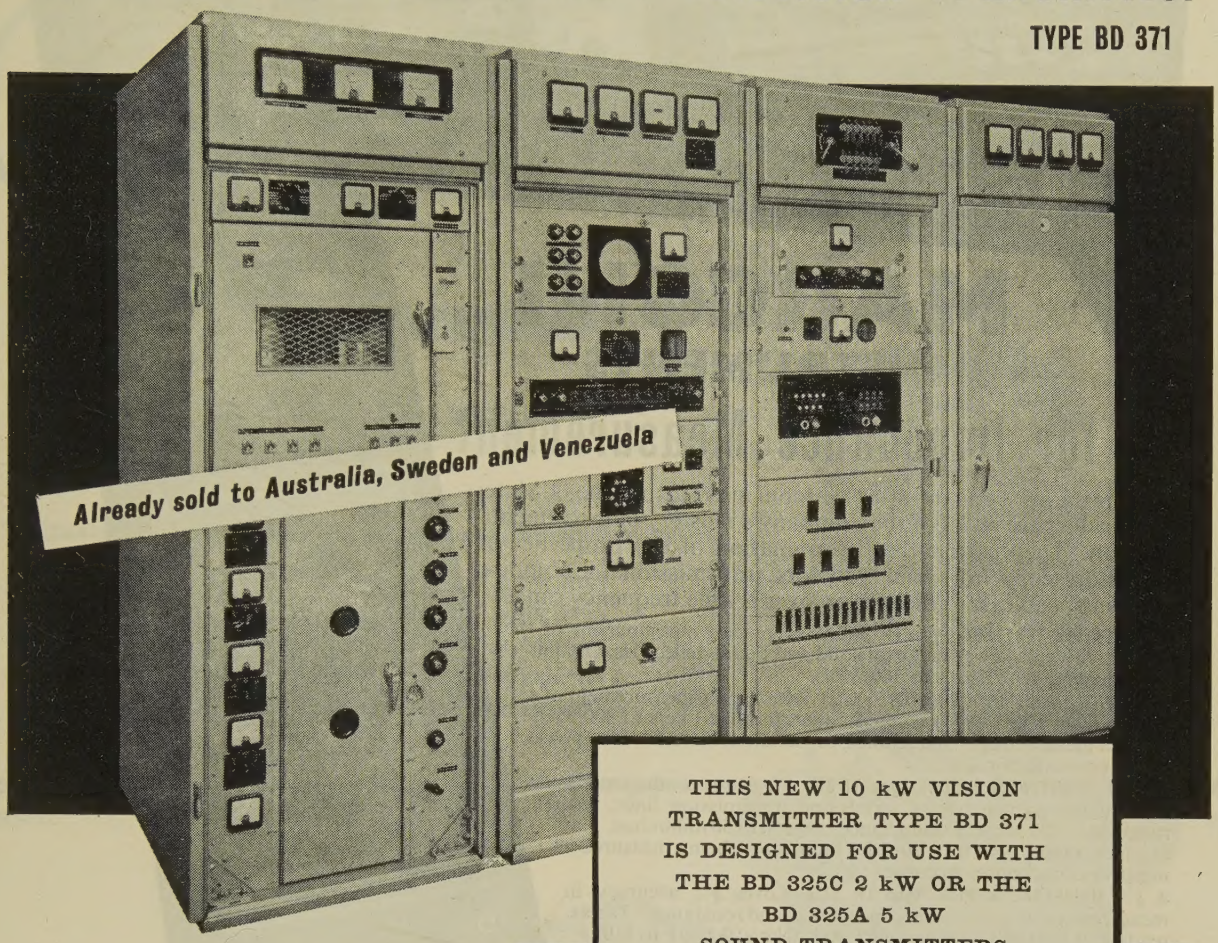
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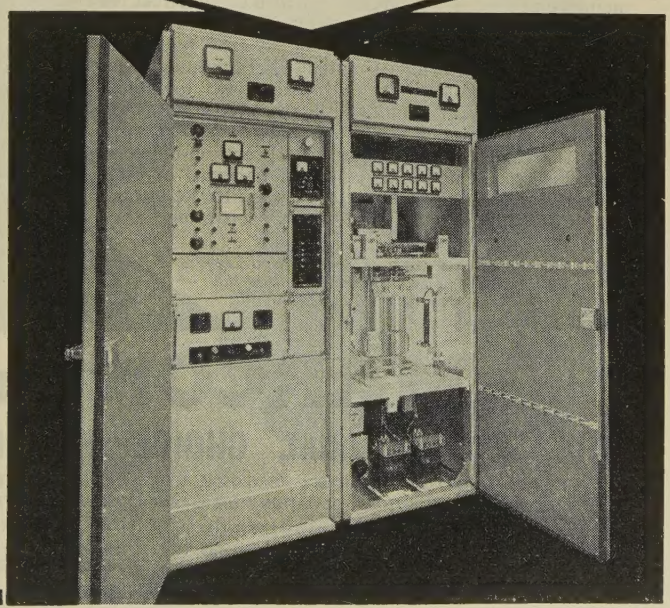
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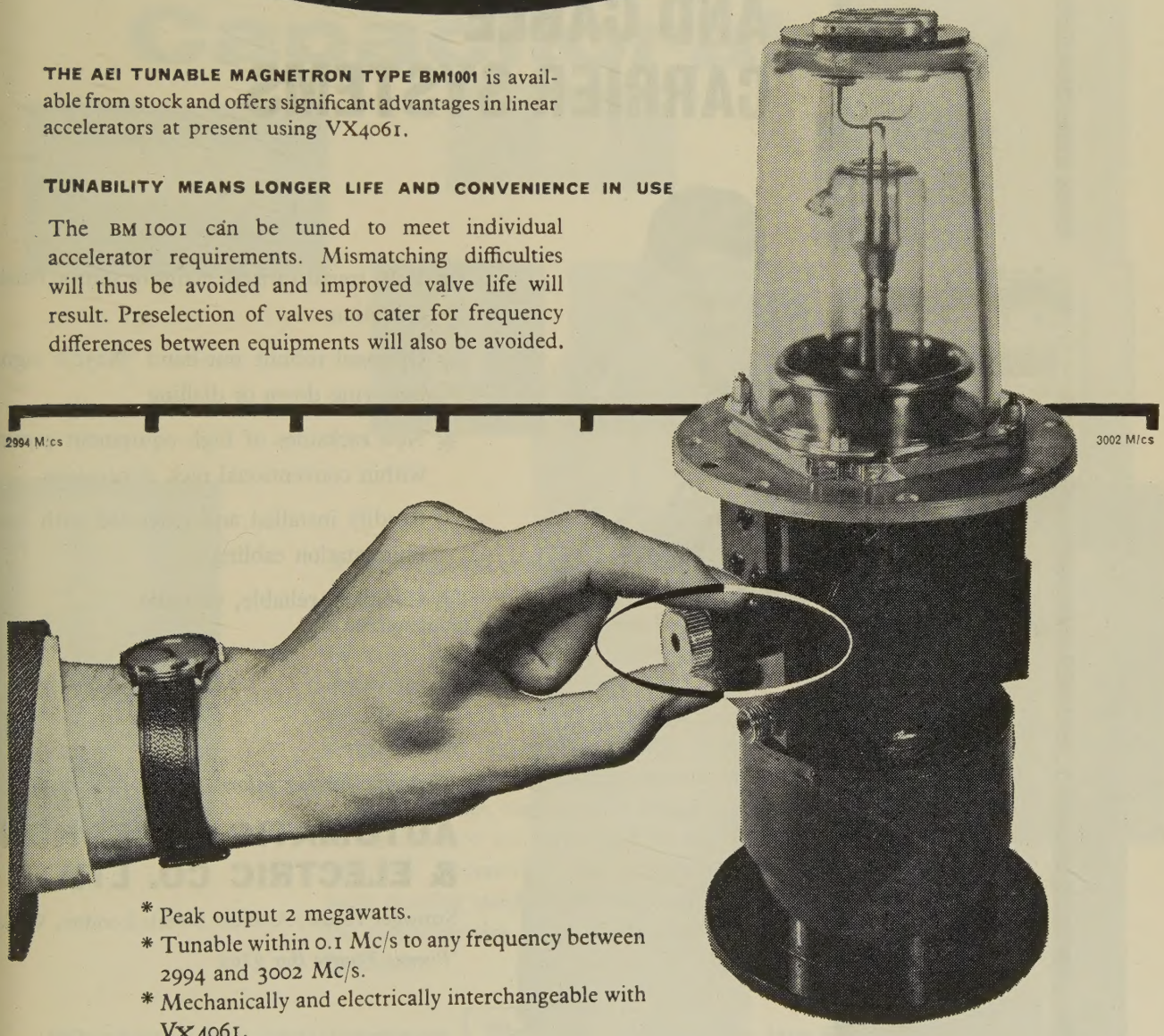
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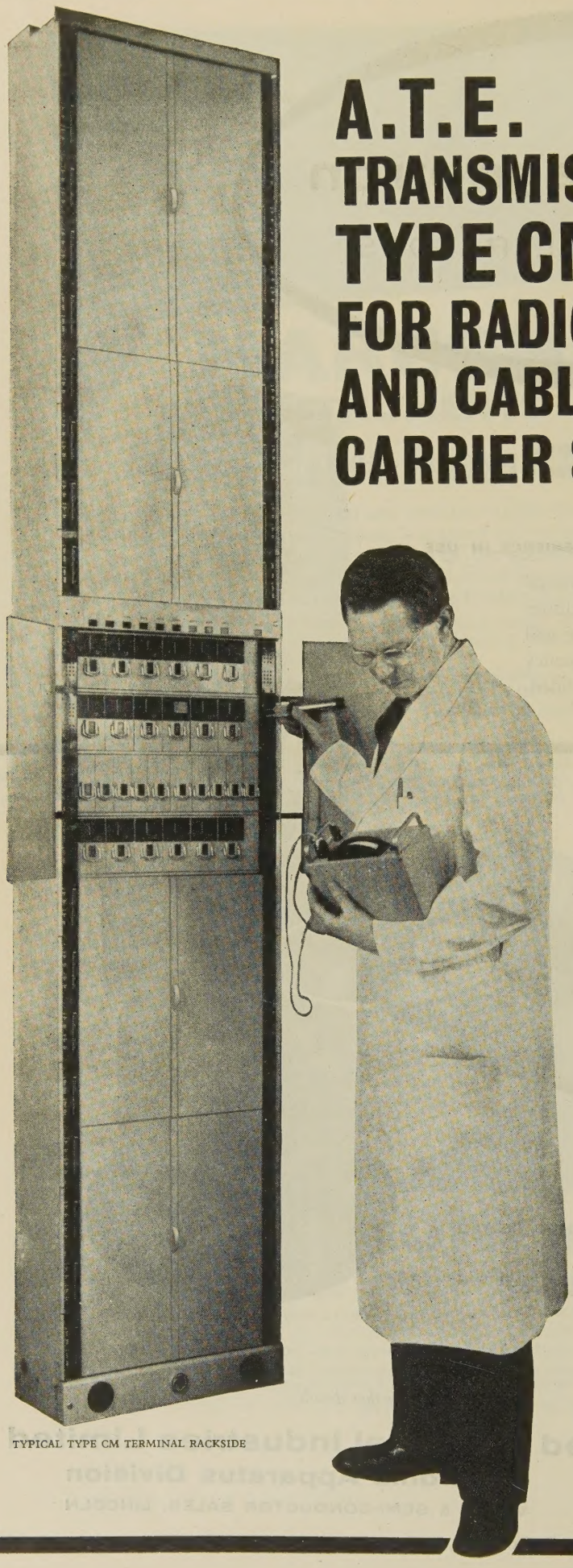
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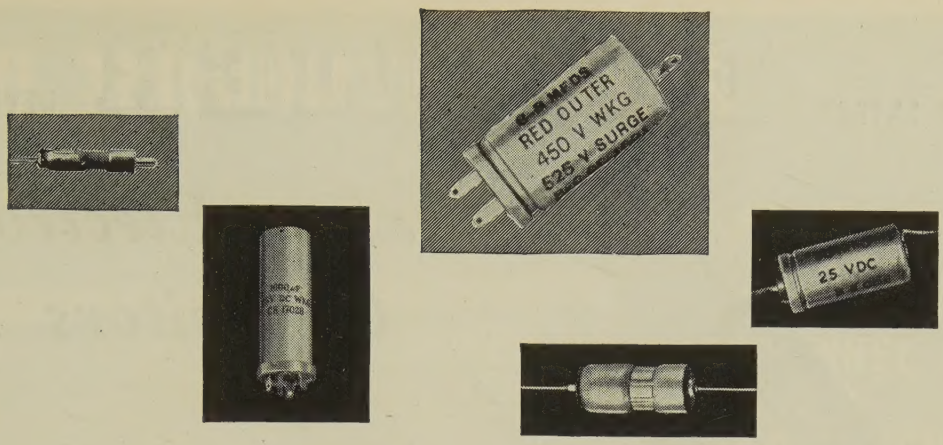
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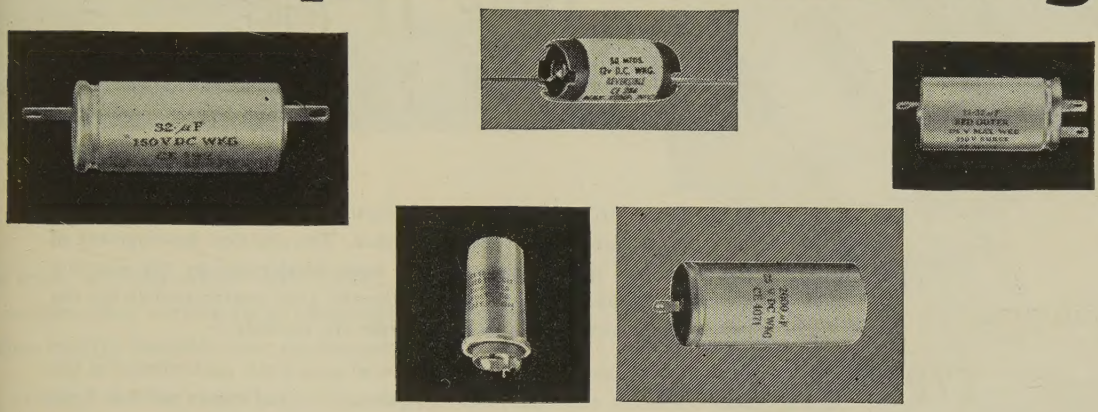
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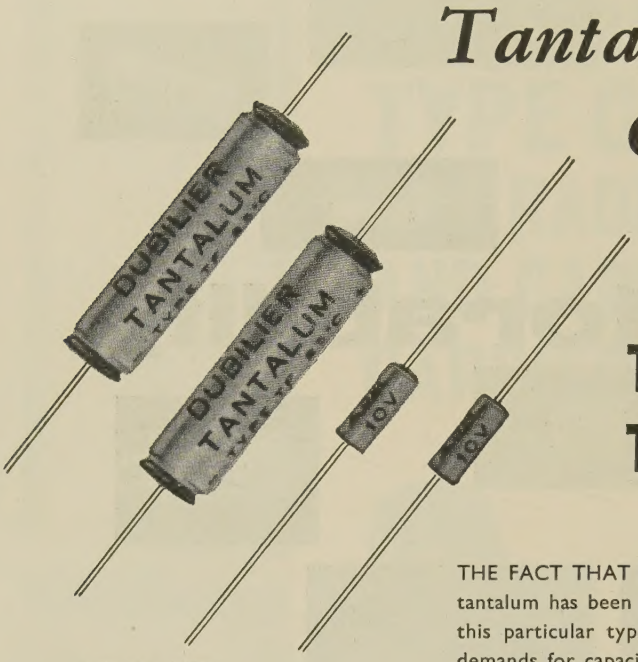
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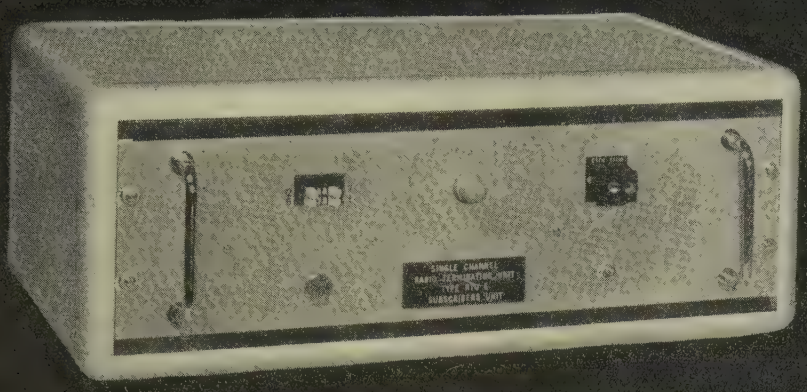
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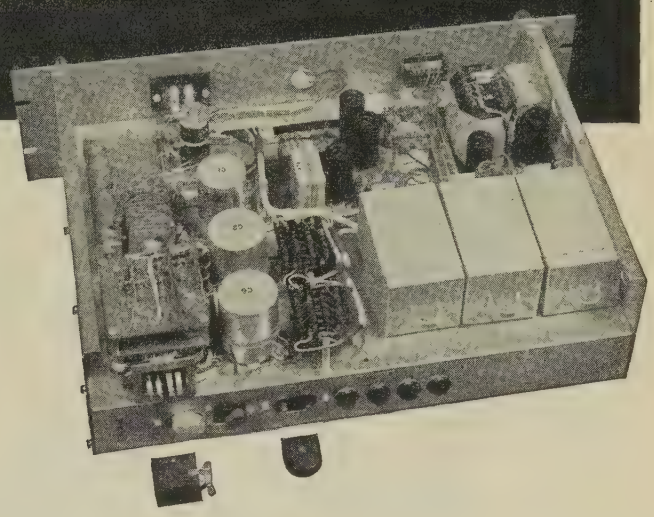
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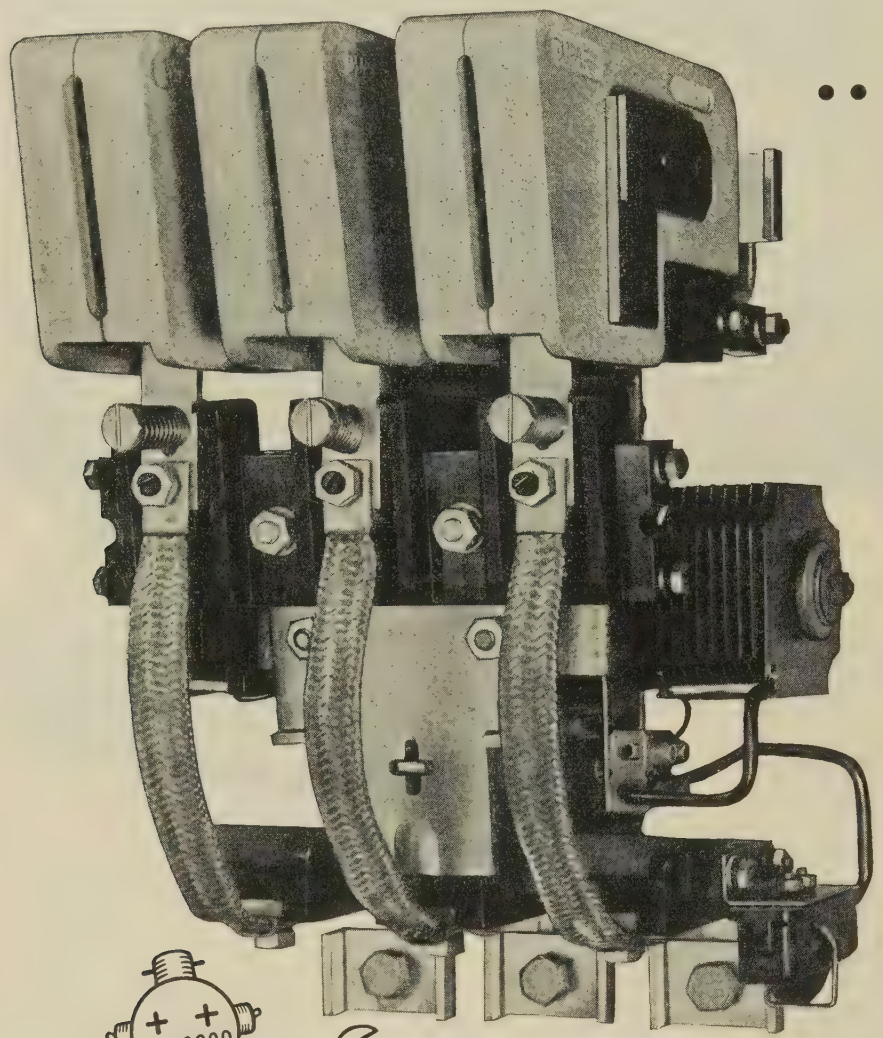
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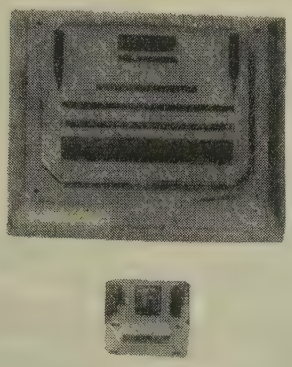
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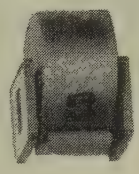
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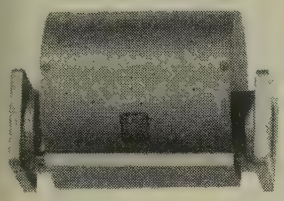
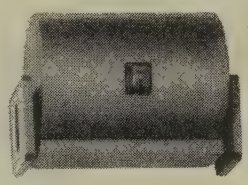
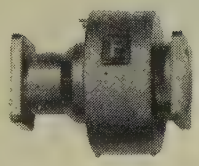
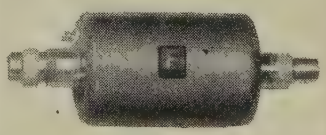
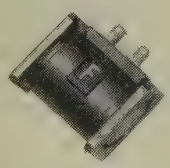
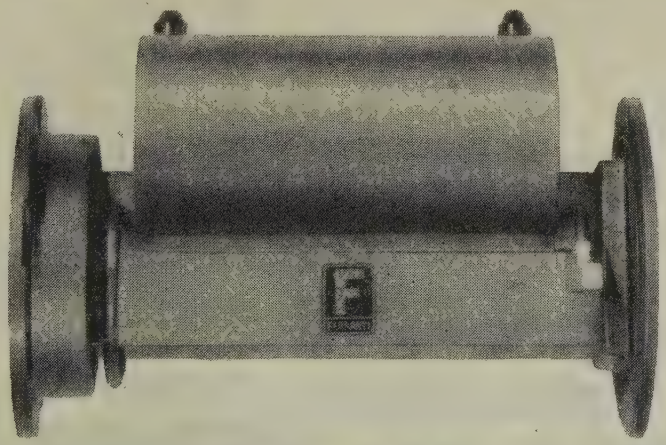
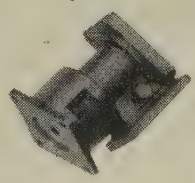
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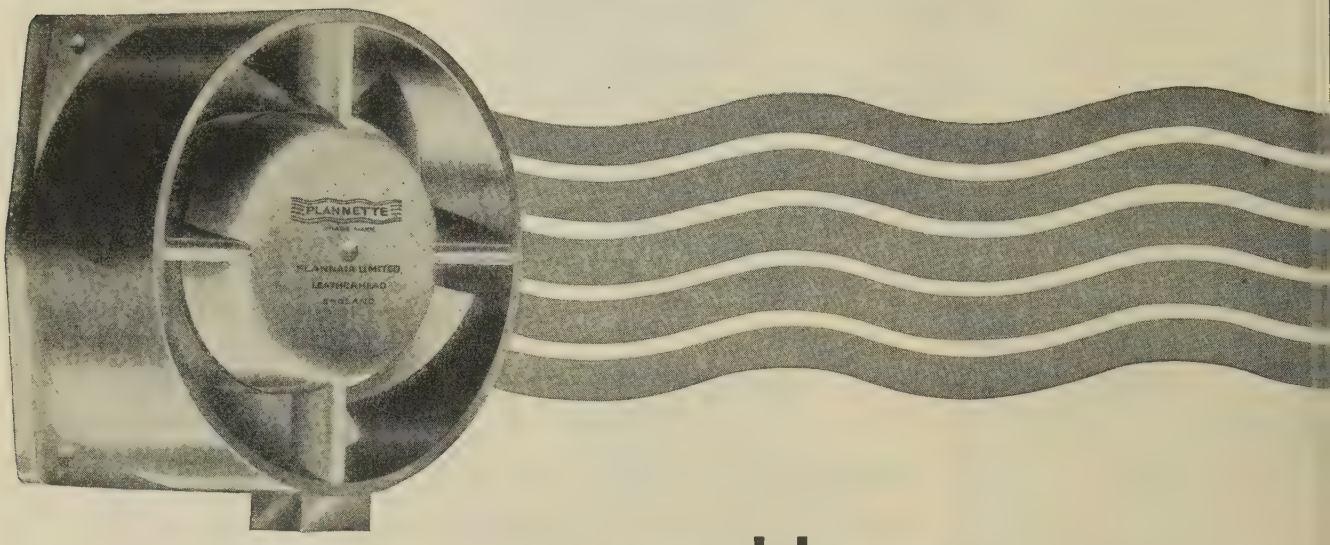
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The new 'Plannette' fans are available in two sizes, 4½″ and 6″ diameter. The casing is light alloy, die cast, and the impeller blades, which are of a special aerofoil section, are moulded plastic.

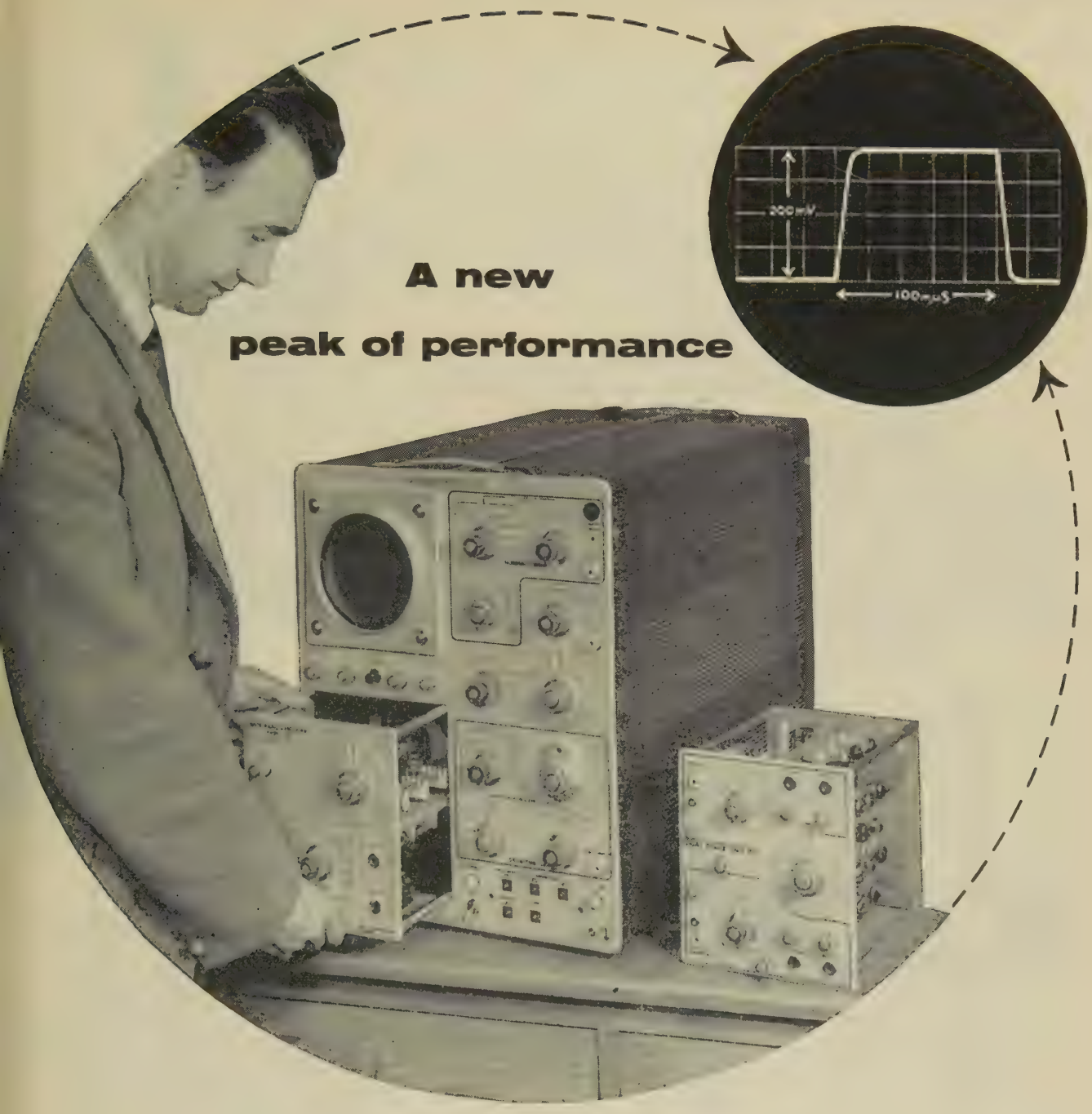
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TIME BASE DELAY	1 μs/–0.15 secs
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Continuously variable from 0·9μsec to 1·05sec i.e. 0·95c/s to 1·1Mc/s. Accuracy ±5%.

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40mμsec. 8V peak in 75Ω, positive going.

Main pulse

Width: Variable from 0·09μsec to 105msec ±5%.

Amplitude: Control gives 4:1 attenuation of each of four maximum outputs as follows:

5V max in	75Ω	rise time 10mμsec
10V max in	150Ω	rise time <20mμsec
25V max in	600Ω	rise time <40mμsec
50V max in	1000Ω	rise time 50mμsec

Polarity: Positive or negative going.

Accuracy: ±2%.

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Obtained from short circuited pure line.
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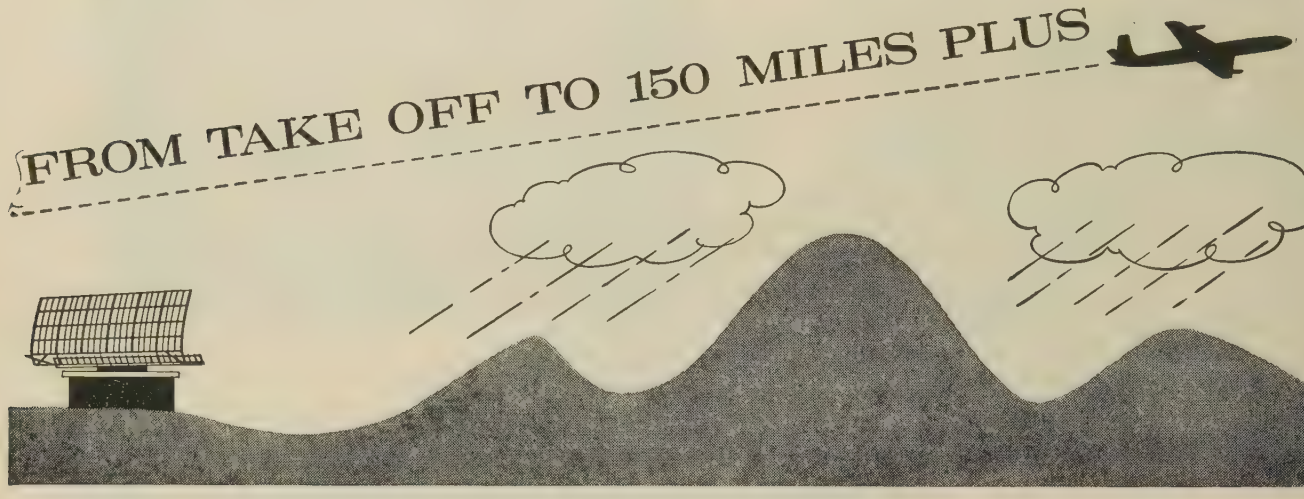
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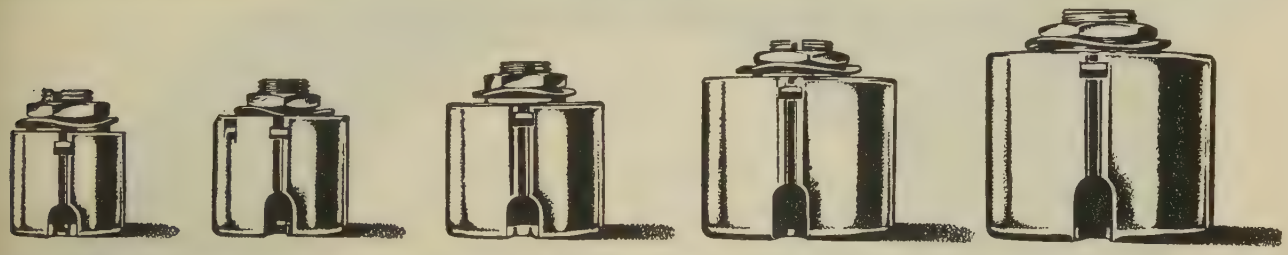
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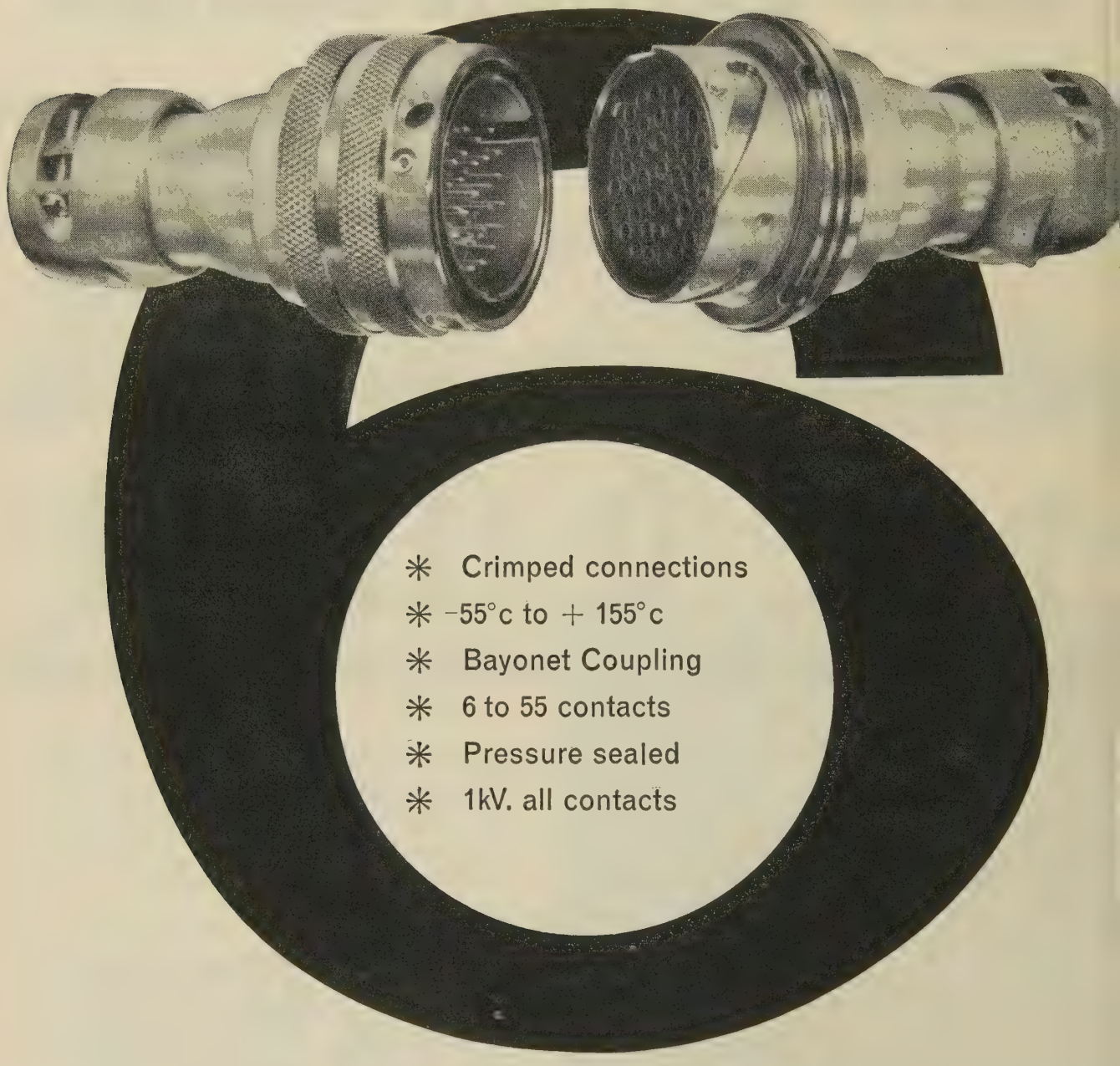
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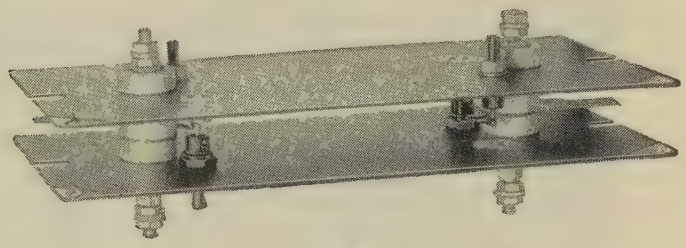
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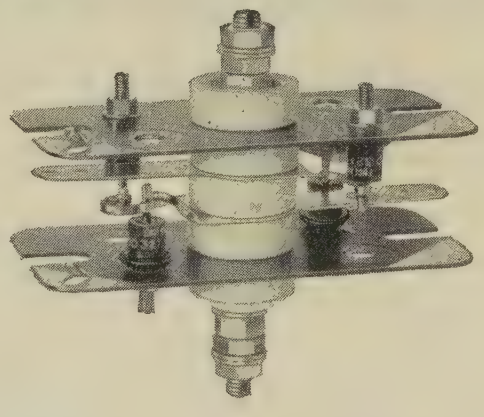
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IB326A	1½×3×2¾	400	280	2·0	150
IB320B	3 ×6×2¾	50	35	2·0	150
IB321B	3 ×6×2¾	100	70	2·0	150
IB323B	3 ×6×2¾	200	140	2·0	150
IB326B	3 ×6×2¾	400	280	2·0	150
IB510A	1½×3×2¾	50	35	10·0	100
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IB513A	1½×3×2¾	200	140	10·0	100
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IB511B	3 ×6×2¾	100	70	18·5	100
IB513B	3 ×6×2¾	200	140	18·5	100
IB516B	3 ×6×2¾	400	280	18·5	100
IB510C	4 ×8×2¾	50	35	20·0	100
IB511C	4 ×8×2¾	100	70	20·0	100
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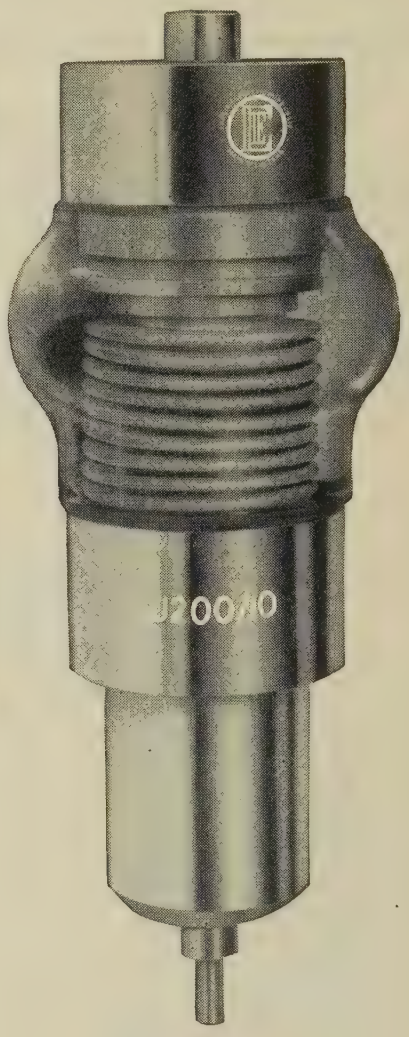
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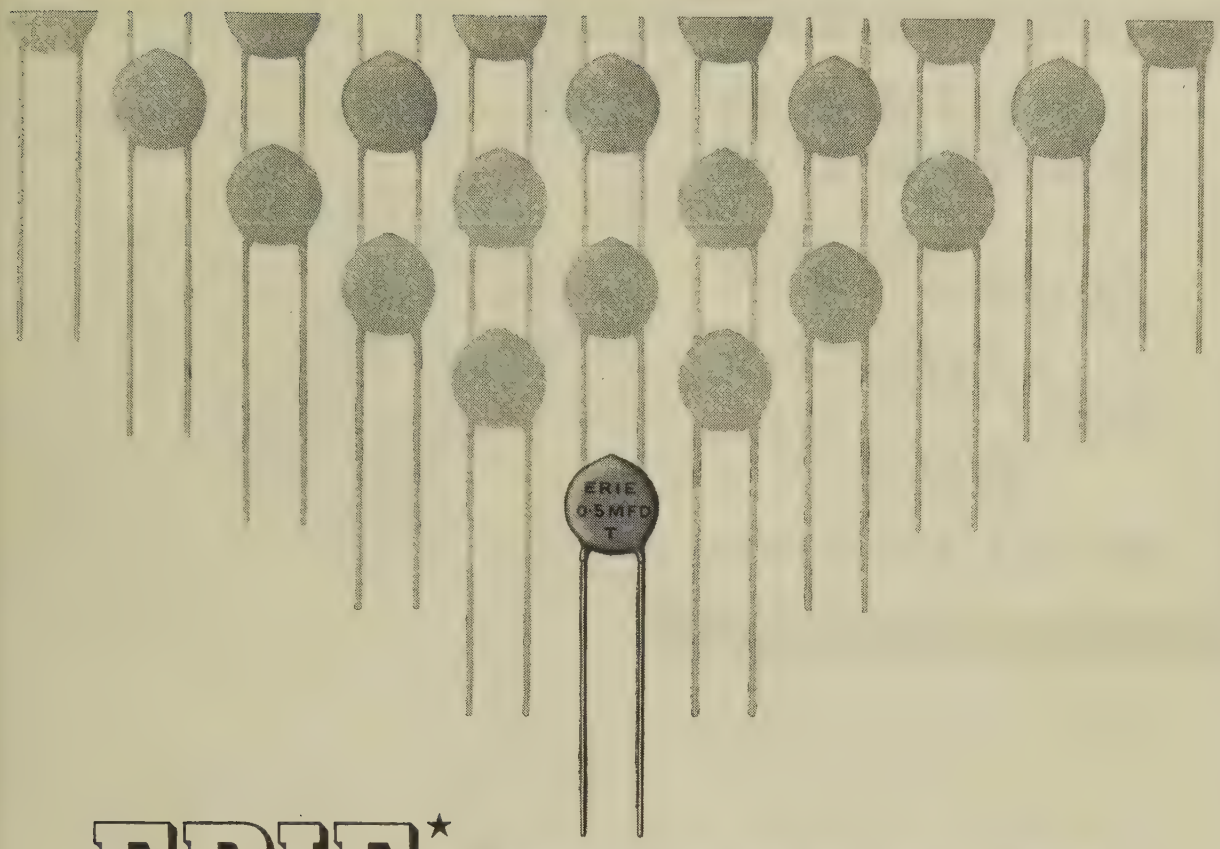
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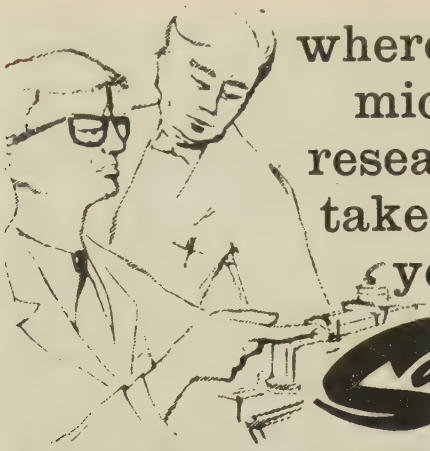
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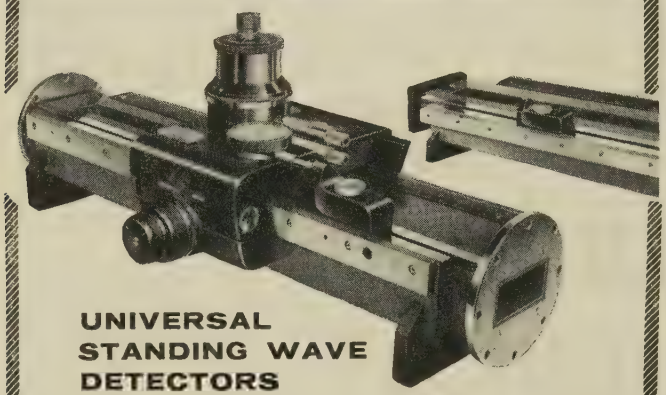
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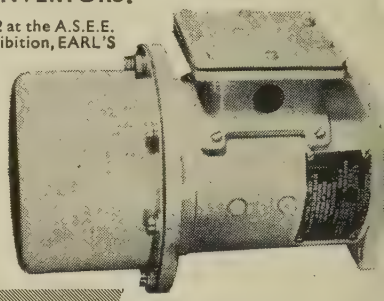
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115 V. 50 c/s



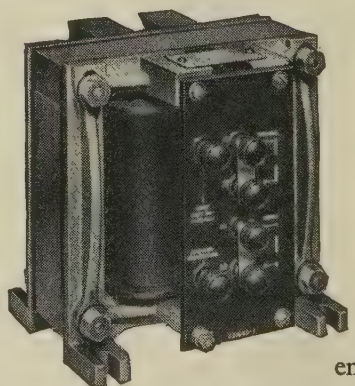
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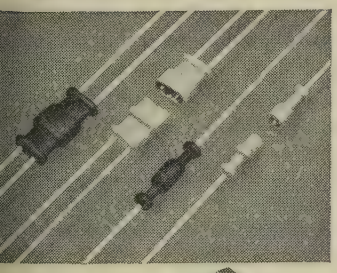
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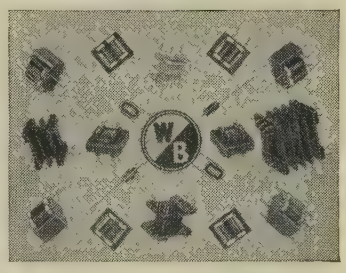


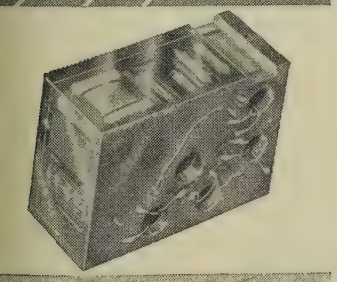
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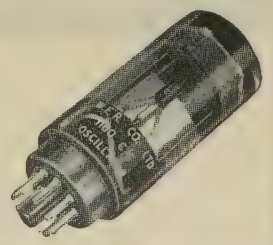


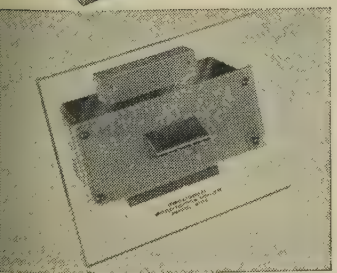
P.O. TRANSISTOR AMPLIFIER

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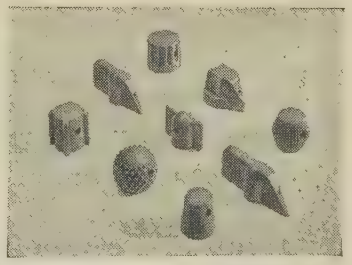


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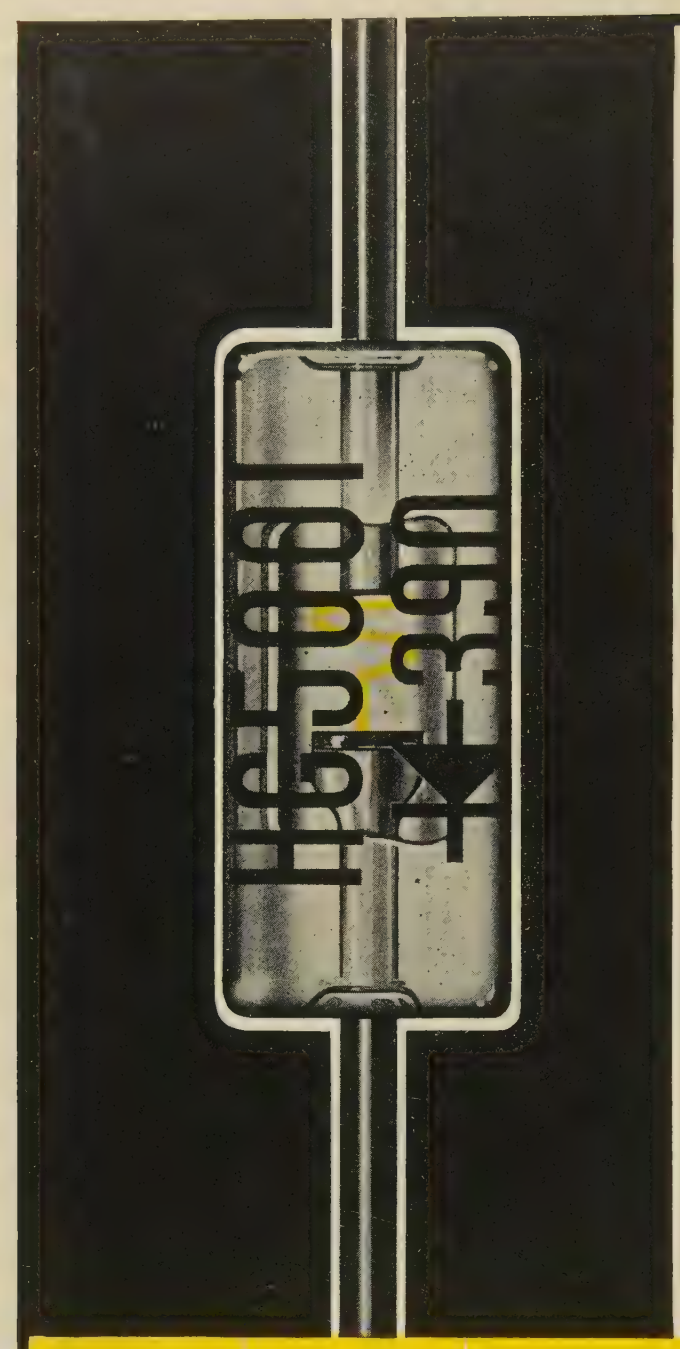
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HG5002	70	80 mA	.8 v @ 100mA	5μA @—50V
HG5003	100	80 mA	.8 v @ 100mA	25μA @—50V
HG5004	70	80 mA	.8 v @ 100mA	25μA @—50V
HG5005	100	80 mA	.8 v @ 100mA	50μA @—50V
HG5006	70	80 mA	.8 v @ 100mA	50μA @—50V
HG5007	40	80 mA	.8 v @ 100mA	5μA @—30V
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L. 108. PART B. NO. 38. MARCH 1961

CENTRE AND GROUP CHAIRMEN'S ADDRESSES

The Institution of Electrical Engineers
Abstract No. 3524
Mar. 1961
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EAST MIDLAND CENTRE: CHAIRMAN'S ADDRESS

By Lieutenant-Colonel W. E. GILL, T.D., Member.

'TELECOMMUNICATIONS—SOME MODERN DEVELOPMENTS'

(ABSTRACT *of Address delivered at* LOUGHBOROUGH *4th October*, 1960.)

From a great variety of changes of equipment and practice, or two examples will be given for each of the principal entials of modern telephony, i.e. the telephone instrument, local line, the switching centre, the trunk network and the rnational link. In dealing with these items, the emphasis be on description as distinct from technicality.

The Telephone Instrument

The two essential components of any telephone system, mainly carbon-granule transmitter and the electromagnetic receiver, fundamentally the same to-day as they were in 1880, and the nciples upon which they work seem unlikely to change in the seeable future. Exhaustive study down the years has, how-r, secured enormous improvements, technically and aesthe-lly, among which may be mentioned increased sensitivity h less distortion, reduction in weight, a wide range of colours d a decreased maintenance liability.

The advantage of greater sensitivity does not lie in louder ech reproduction than that to which the user is at present ustomed; this in itself, as will be seen, could be an embarrass-nt. It lies in the ability to obtain adequate speech volume r longer lines and in the great economies to be secured by lizing lighter-gauge conductors on shorter lines.

The source of power for a telephone transmitter is generally ived from the exchange batteries, and the current varies between the short and long lines. It is necessary to keep such iations within limits, and hitherto this has been done by ans of a barretter, which automatically adjusts its resistance suit the length of line. With the latest instruments, however, arretter alone cannot compensate for the wide range of line istance over which they will operate, and to avoid, on short es, excessive loudness and possible acoustic shock, a line-rent sensitivity regulator is included within the telephone trument itself. Basically, this automatic regulator is a variable-s network made up from rectifier elements inserted in series h the transmitter.

Adequate speech volume is only one of a number of requirements to be satisfied, however; an equally important one is freedom from distortion. Very high-quality reproduction is practicable at a cost, but it would be quite uneconomic to provide anything better than that which ensures that speech is easily understood. For good-quality speech reproduction a telephone must transmit and receive over a frequency band of 300–3 400 c/s, which is, in fact, an international standard. Any increase in range above 3 400c/s would mean increased costs without a proportionate increase in intelligibility, and therefore the latest instruments do not extend the frequency range but achieve a more uniform flat response which results in reduced distortion.

Although the new instrument has a 'new look' in shape, weight, colour, cordage, printed as well as conventional wiring, gravity switch, ease of maintenance, the only additional item is the automatic regulator and the only great innovation is the use of a rocking-armature receiver in place of the magnetic diaphragm. The great increase in sensitivity is due almost entirely to the rocking-armature receiver, which, in effect, separates the magnetic and acoustical functions of the receiver itself.

The Local Line Network

Another essential item in the subscriber–exchange link is the local line, and with certain exceptions each subscriber requires an individual pair of wires directly connected to the exchange. Planning provides for bulk provision of these pairs by underground cable to selected distribution points in an exchange area and thence by short overhead or underground wires to the subscribers' premises.

Ideally there will always be a spare pair of wires available in underground cable with only minor overhead construction required to effect a new connection. This ideal is achieved only if demand for service comes where expected and at the rate forecast. This does not always occur and can lead to either under- or over-provision, which in turn involves capital expenditure or capital lying idle. This difficulty has been much reduced

Lt.-Col. Gill is with the General Post Office.

1961: The Institution of Electrical Engineers.

by the introduction of a cabinet-pillar distribution-point flexibility scheme for the distribution network which readily permits diversion of cable pairs between terminal points.

Essentially a *cabinet* contains a cross-connecting frame on one side of which is terminated the main cable pairs from the exchange. On the other side are terminated cables radiating out to a number of *pillars* and the cross-connection facilities in the cabinet permit selected pairs from the exchange to be put through to these pillars. Similarly, the pillars themselves are cross-connecting points permitting the use of any pair from the cabinet to be connected to cables radiating out to a number of distribution points.

By judicious choice of cable sizes, maximum flexibility is obtained and fluctuation in demand is catered for more economically and efficiently. Economically because of reduction in the number of pairs required to be provided in the main cable—the most expensive part of the network—and because diversions of pairs at cabinets and pillars is cheap compared to diversion in cable joints. Efficiently because of speed and the removal of risk of damage in opening up the larger cable joints.

The drive for economy in line construction is illustrated again by the introduction of lighter-weight copper conductors for cables. A 4 lb/mile copper conductor is now in general use, whereas the minimum previously employed was 6½ lb, and it is anticipated that with improved jointing technique even lesser weights will be available in the not too distant future. This has been made possible by advances in equipment design, particularly the new telephone instruments, already described, which have increased both the permissible d.c. signalling and transmission loop resistances of a subscriber's circuit from 650 to 1 000 ohms.

The economies to be obtained will generally be obvious, not least, with the increased length of line now obtainable, the possibility at some future date of a reduction in the number of exchanges.

Some experimental work has also been carried out with light-weight aluminium conductors for cables, but until such time as a much simpler and improved jointing technique, both aluminium to aluminium and aluminium to copper, is available, copper seems unlikely to be displaced. The future of copper as the conductor is in fact reasonably assured, but that of lead as the sheathing medium is not so. The disadvantages of lead for sheathing are weight and susceptibility to corrosion and mechanical damage, particularly the last two when the cable is laid direct in the ground.

Underground cables are in the main accommodated in single or multi-way self-aligning earthenware ducts, a method which facilitates the laying of cables and of replacement of existing ones. This method is, however, very costly, and there are many situations where a duct line is an unnecessary luxury. As stated, however, experience with lead-sheathed cables laid in this manner has not been at all satisfactory and special protective measures in the event can be as costly as provision of a duct line.

The advent of plastics of the polythene and p.v.c. types pointed the way to a solution of this particular problem, with the result that lead is now being rapidly superseded as the material for the sheath of telephone cable. Furthermore, polythene has such excellent electrical properties that it is also replacing paper as the insulant for the wires in the cable. To date all local cables up to 100 pairs are of the all-polythene type, and the range will undoubtedly be extended. External cabling is the largest factor in the cost of providing a subscriber's telephone and offers the greatest possibility of savings. The use of polythene is a most important measure towards realizing such savings.

Polythene cables in the present range are appreciably less costly than corresponding lead-covered cables; there is little difference in respective diameters; they are lighter, tougher more easily handled; they do not corrode readily; they ca drawn into ducts in lengths up to 1 000 yd or laid direct in ground; and they are standard for use as aerial cables. T are, of course, disadvantages, the principal one being the f cation of the necessary joints, and this, until such time a improved moulded joint is made available, remains a diff operation.

The present method employs an expanding-plug joint addition, because open-ended sleeves are employed for actual conductor joints, it is essential to prevent any water w may find its way through the cable sheath from entering actual joint, and this necessitates the fitting of water ba on either side—a tedious, rather difficult operation for whi special wax is employed.

Regarding external plant, one radical change of practi worth mentioning and concerns the protection arrangem for telephone subscribers' circuits against lightning and o extraneous voltages. With the exception of certain modificat the standard method of protection by fuses, heat coils and ca block mica-strip spark-gap lightning protectors fitted at the exchange and subscribers' termination has remained virtu unchanged since early in this century. Full protection, u practice to date, is still maintained at the exchange becau segregation of completely underground circuits from t which include overhead wires is a practical difficulty. subscribers' premises, first heat coils were dispensed with enti and later protectors were omitted on wholly undergro circuits or with circuits of four spans or less of open wire. change of practice involves change of location of the light protector at present fitted at the exchange termination.

Research into lightning damage to cables has shown th great proportion of this occurs at the distribution point o the cable leading to it. Such damage not only causes se interruption to service but is expensive to repair, and it fol that the place to give protection is at the immediate junctio the overhead and underground network, i.e. at the distribu point.

This method, called pole-top protection, involves the inse of an electrode protector in each wire of a circuit at the d bution point. For routes of less than six pairs, these prote are fitted into the existing terminal insulator, but for la routes they are built up into units and mounted directly on poles, a suitable earth lead being provided. These units re the existing block terminals.

At the exchange termination a new fuse of the delay-ac type is provided which, in effect, combines the functions o present fuse and heat coil to protect the equipment agains effect of both small and large continuous extraneous curr Again, the economies to be obtained not only in mainten costs, but in saving in rack space in an exchange, will be obv

The Switching Centre, or Telephone Exchange

Amongst the many outstanding developments in this field are of particular interest: the move towards full autom and the imminent advent of the all-electronic exchange.

Dealing first with the electronic exchange, research and de field trials have advanced to the stage where some electr directors or register-translators have been in service for a nu of years and are giving eminently satisfactory results. Also experimental electronic exchange is being exhaustively teste the Dollis Hill Research Station, which, although not a p exchange, has access to the London telephone exchange netw There is every probability that the first public exchange of type will be installed in London within the next two years.

The employment of electronics in exchange design will result smaller and less equipment, less building space, greater iability because of the virtual absence of moving parts, and imately reduction in initial costs.

Turning now to full automatization of the telephone system, e first move was taken in the 1920s with the decision to adopt e Strowger step-by-step system, and the final one came with e fairly recent decision to give subscriber trunk dialling ilities throughout the United Kingdom within a period of n years.

It should be noted that the economies of telephone growth and age are the important factors dictating this decision. In fact, e telephone service has expanded from some 1 300 exchanges d half a million stations in 1906 to over 6 000 exchanges and million working stations in 1960, of which 75% already have tomatic facilities. The size of the problem is perhaps best dicated by the fact that in 1959 some 340 million inland trunk lls were made in the United Kingdom.

Subscriber trunk dialling means precisely what it says—that y one subscriber in the United Kingdom can obtain by code alling any other subscriber in the United Kingdom. Already e way has been paved to the extent that mechanization of e trunk system has resulted in a large proportion of trunk lls being now set up automatically by the originating manual erator—the irreducible limit under manual conditions—and th the completion of the auto-trunk switching network at an rly date and the installation of the required auto-switching uipment, the way is clear for 100% automatization.

The Auto-Trunk Switching Network

A speech circuit over 25 miles in length is designated a trunk rcuit. The increase in trunk calls already mentioned is, of urse, reflected in the increase of such circuits. There are, at esent, some 25 000 of them serving the public network, com-red with less than 2 000 in 1920.

At the beginning of the century, such relatively few circuits existed employed 800 lb/mile copper wires on the stoutest of les. With the introduction of inductive loading followed by nplifiers, a great reduction in weight of conductors was effected d the use of cable for trunk circuits was made practicable. In 1930, carrier current working was developed for use in ephone cables, and at present it provides 960 circuits on two axial pairs of $\frac{3}{8}$ in diameter with 6-mile spacing of repeater ations. The latest coaxial system at present under construction n cater for approximately 2 000 speech circuits using two n coaxial pairs with a 3-mile spacing of repeaters.

The trunk network is designed to ensure the most efficient e of expensive line plant in terms of transmission and occupancy ne with the maximum economy in capital outlay. It must ter not only for heavy traffic between large cities but also for casional traffic from any one village to any other remote lage.

In the basic layout of the system, traffic-collecting ('marshalling rd') centres are employed. These are termed zone and group ntres, this country being divided into 18 zones, each subdivided o a number of groups. Every exchange in the country has cess to a zone or group centre and therefore access to the main unk network.

Standardization of transmission is achieved by ensuring that zone-to-zone and all zone-to-group speech circuits have actly equal characteristics. It should be stressed this is the sic system which is by no means rigid; but, in general, a group ntre collects the trunk traffic, routes it via its own zone centre, hich is interconnected to every other zone centre, and hence any other required group centre, which in turn distributes it.

The International Telephone Network

Perhaps the most significant advance in the sphere of international communication is the rapid development of the submarine telephone cable. Such cables are no novelty over such short distances as the English and Irish Channels, but up to 1927 the field of long-distance communication had been firmly held by the telegraph since 1866, when the first transatlantic cable was successfully completed.

In 1927, the radio circuit entered into competition and the first commercial telephone link between the new and old worlds was inaugurated between London and New York; and gradually the international radio-telephony network took shape to link the continents.

By the early 1950s the limit on further additions to the long-distance radio-telephony network had virtually been reached because of exhaustion of frequency space, this applying, in particular, between America and Europe. Such radio circuits, however, although giving good service, are sometimes uncommercial because of atmospheric conditions.

A passage in the Construction and Maintenance Contract signed by Lord de la Warr, the Postmaster General in 1953, on behalf of the British Post Office, which reads '. . . And whereas it is desired to provide a submarine telephone cable system between the United States and Canada on the West and the United Kingdom on the East . . .', is at once of historical significance and a milestone in electrical engineering vision and achievement. Known as the T.A.T. and completed in 1956, this project costing some £15 million provided, for the very first time, speech circuits in cable between the two worlds. Thirty-five high-quality telephone channels became available, 29 to New York and 6 to Montreal, in addition to a telegraph channel to Canada and maintenance circuits, to assist in meeting the ever-increasing demand on traffic capacity.

Technical progress in the design and construction of submerged repeaters, together with great improvement in mechanical design of the structure of the cable itself and application of the technique of carrier working over coaxial paths, made this achievement possible. Apart from the complex links connecting the shore ends of the cables to the switching centre, it is a measure of the task and of the confidence of all concerned that the project involved the manufacture and laying of 4 500 miles of submarine cable and 146 submerged repeaters, built to withstand laying hazards and water pressures at depths exceeding $2\frac{1}{2}$ miles and whose components are required to have at least a 20-year life without attention. This is perhaps the most critical of all the applications for which electronic components are used.

Briefly, the deep-sea cable link consists of two parallel single coaxial cables each incorporating 51 repeaters, the shore terminals being at Oban, Scotland, and Clarenville, Newfoundland. Each repeater is equipped with some 300 electronic valves and 6 000 other components; and the repeaters are energized from shore-end-fed direct-current supply over the centre coaxial conductors at voltages up to 6 kV.

Such was the immediate commercial success of this cable that in September, 1959, a second cable, designated T.A.T. 2, was opened for service. This links New York with Paris and Frankfurt, with a deep-sea link terminating at Clarenville, Newfoundland, and Penmarch, France.

Yet another transatlantic system, the T.A.T. 3, is at present in the planning stage, for completion in 1963, providing an additional 128 circuits between the United Kingdom and the United States. Another such project of immense importance is the Commonwealth round-the-world telephone cable, construction of which has already commenced. It is over some 32 000 miles in length, and will cost £88 million. The first section, known as Cantat, will be in service in 1961, with a 60-circuit capacity

linking directly London with Quebec and Montreal. This will be followed by project Compac between Canada, Australia and New Zealand.

Cantat is a radical new type of deep-sea cable—the 'light-weight'—an unarmoured cable carrying the strength member in the form of a high tensile steel strand at the centre, within the inner conductor of the coaxial pair. The ubiquitous polythene is utilized for the overall sheathing, and the overall diameter is exactly the same as the armoured type used for T.A.T. 1 and 2. The circuit capacity is 70% greater than that of the T.A.T. systems; both 'go' and 'return' channels are carried in the same cable, so that only one cable is required instead of two, with fewer repeaters. This is a much cheaper system to install and maintain. However, because of the risk of damage to such a cable at shallower depths, armoured cable will continue to be used up to 500 fathoms.

This particular aspect cannot be left without brief mention of time assignment and speech interpolation. In any telephone conversation over a 4-wire 'go'-'return' channel the complete circuit is occupied for only half the time, and by the addition of suitable equipment the circuit capacity could theoretical doubled. A high-speed selection and switching device, d nated time assignment and speech interpolation, has produced and has already resulted in increasing the ci capacity of the existing T.A.T. 1 and 2 systems.

I have touched briefly and in non-technical language on a of the numerous interesting and far-reaching development telephony; hardly a day passes without news of yet more.

The scientists operating Jodrell Bank announced, al casually, in 1959, their success in transmitting both teleg and speech messages from there to America via the moo half-million-mile link.

The very thrust of economic development of the world at with its ever-increasing populations creates a continuous growing demand for rapid, efficient and inexpensive comm cations of all types. Telecommunication is essential to development and growth, and this branch of electrical engine must always be ahead of, and not just content to keep up the pace of the modern world.

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SOUTH MIDLAND CENTRE: CHAIRMAN'S ADDRESS

By Brigadier F. JONES, C.B.E., M.Sc., Member.

'THE GROWTH OF THE TELEPHONE SERVICE'

(ABSTRACT *of Address delivered at* BIRMINGHAM 3*rd October*, 1960.)

The telephone service, like other branches of electrical engineering, owes its growth to the increasing pace of technical development which has taken place in the last half-century, and like them it has had profound effects in industry, in commerce and in our daily life. In offering a direct and immediate means of communication by word of mouth rather than by transmission of a written message, however, it has altered the outlook of the individual in a more fundamental way. Thus the growth of the telephone service at first depended mainly on the readiness of the public to accept this change in outlook, and up to the early years of this century the growth of the telephone habit tended to be slow in Europe but much more rapid in North America. In 1910, the number of telephones in use in the United States was about 8 million, but was less than half a million in this country.

Soon after this date, however, the era of rapid technical advance began and two developments took place which opened the door to an almost unlimited extension in the range and scope of telephone communication. The thermionic-valve amplifier began to be used in the cable network, and soon after the First World War a systematic programme of converting local manual exchanges to automatic working was inaugurated. As the valve amplifier extended the possibilities of trunk communication, so the automation of telephone exchanges removed the limitations imposed by manual operation on the size of the system. The steady progress made in the following decade in these applications resulted in an immense increase in capacity. The replacement of overhead routes by repeatered cable was later supplemented by multi-channel carrier systems, and just before the last war some progress was made with coaxial cables. At the same time, automation of local exchanges permitted an extension of capacity free from manpower restrictions. Under these circumstances, the facilities for growth soon began to out demand, and in the early 1930's efforts were made with siderable success to stimulate demand by the incentive of red rentals and call charges.

The outbreak of war in 1939 put a stop to all growth as fa the public service was concerned. Demand suppressed du the war therefore accumulated to be released in a surge w followed immediately afterwards. It has been met by a m increased rate of installation, and the number of subscribers more than doubled since the end of the war. Even now have not yet everywhere achieved the position of being abl provide service for all our customers. Progress in the tas overtaking these arrears and in providing plant and equipm on such a scale that service can be given to anyone immedia on demand has been handicapped by limited capital resou The greater part of capital expenditure on the telephone se is taken up by external cable plant, and various expedients been introduced to make the fullest use of it. The most im tant in its effect on the public has been the adoption of sh service, which, though not universally popular in this cou has succeeded in giving service to many who could not other have had it. Experience has shown this form of service t quite acceptable by the great majority of sharing subscribers total number of whom exceeds one million.

The proportion of the time during which the average reside subscriber's line is in use is very small, and another de known as the 'line connector' is now being developed, w enables the high proportion of free time during which the is unoccupied to be exploited more fully. The idea is to ex the common-user principle on which exchange equipmen provided to include the subscriber's line by the insertion in local distribution system of switching equipment which ena any one of a number of subscribers to be connected to a line to the exchange when making a call. In this way

Brigadier Jones is with the General Post Office.

scriber has the exclusive use of the exchange line during his , and a grade of service can be given equivalent to that of the hange as a whole while making the most effective use of the plant.

he rapid increase in the number of subscribers has not been tched by a corresponding increase in telephone traffic, and calling rate, i.e. the number of calls per telephone, has shown arked decline over the last forty years. The explanation lies he increasing proportion of residential lines, which are used eneral much less than business lines; and to stimulate greater of the residential telephone the cost of local calls has ntly been reduced. Trunk traffic, on the other hand, shows se more in proportion to the increase in the size of the system has been rising at an average rate of about 5% per annum; r the last two years the annual rise has been 10%.

n order to cater for this growth the network has been nforced mainly by the construction of coaxial-cable systems ing the main centres. Each coaxial-cable tube is capable transmitting a minimum of 600 telephone channels or one vision channel, and on some recent systems, more frequent cing of repeaters has enabled many more channels per tube e derived.

n the last few years, radio links, operating in frequency ds up to 30 Gc/s, have been used increasingly to augment trunk network, and a main north–south system of radio s is being planned which will provide several thousand phone channels.

Much work is now being done on the use of circular waveguides long-distance transmission, and the possible development of tems which will allow frequency bands of the order of Gc/s to be used for commercial transmission opens up a spect of increasing the number of available channels to ny times that provided by the systems at present in use.

While the long-term process of converting the local telephone vice to automatic working is still proceeding, the introduction automation to the trunk service has been undertaken in the 10 years. It has followed two stages. The first stage was installation of automatic equipment at trunk centres, so that dialling through this equipment the controlling operator can up the trunk call directly and without the assistance of any er operator.

he second stage for complete automation requires the replace- nt of this remaining operator by apparatus which is capable of rying out automatically the functions of routing the call to required number and applying the appropriate charge for the according to its distance and duration. Trunk dialling by scribers has already been introduced at some exchanges, and present programme provides for its extension over most of country within the next 10 years.

Under subscriber trunk dialling a method known as periodic ering is used for the automatic recording of timed calls, and subscriber's meter, hitherto only used in registering local calls, mployed for trunk calls as well. The unit charge for trunk ocal calls is the same, namely 2d., but the time allowed varies ording to distance. Thus, over the maximum distance the e allowed for the unit fee is only 12 sec, while for shorter ances the time interval is progressively longer up to a period min for a local call.

he obvious advantages of subscriber trunk dialling are speed, hat the subscriber can dial direct instead of seeking the ices of an operator, and cheapness, in that the effect of the system of charging is that a short-duration call over the gest distance can be had at a very much reduced rate. Present mates of the rise in trunk traffic attributable to subscriber k dialling at those exchanges where it has already been oduced indicate a rise of 40%.

In recent years many developments have taken place in the field of electronics. One of these has been the application of electronic means of switching in telephone exchanges in place of the electro-mechanical method on which the present system in this country is based. The elimination of mechanical moving parts, with their inherent liability to wear and tear, and their replacement by electronic impulses offers obvious advantages, one of the most important of which lies in rapidity of operation. Quicker operation means a saving in equipment, with corresponding savings in accommodation and maintenance.

The rapidity of electronic switching can be exploited to enable a number of telephone calls to be transmitted over a single channel by allotting to each call in turn a very short period of access to the channel. The system, which is known as time-division multiplex, is employed in the design of an electronic exchange soon to be opened at Highgate Wood in North London. The principle is not new and has been applied in the past in telegraphy, but for the purposes of telephone transmission a frequency of switching is required which can only be achieved by electronic methods. Thus, if a subscriber's line is connected to a transmission channel for a period of 1 microsec, and if the connection is repeated 10 000 times per second, speech can be transmitted over the channel without apparent discontinuity or distortion. Since the subscriber is only connected for 1 microsec in 100 microsec, it is possible to transmit 100 calls simultaneously over the same channel.

So much progress has been made in the application of electronics to the telephone service that the first subscriber-trunk-dialling installation at Bristol incorporated routing and charging equipment of electronic design.

An even greater saving in space requirements is achieved by a later design of equipment employing a magnetic-drum technique. In this type the information can be recorded and stored on the surface of the drum by the formation of minute magnetic cells. Because of its small space requirements, this equipment is particularly suitable for centres where heavy traffic is to be handled.

Automation is also being applied to the preparation of subscribers' accounts. The manpower involved in this work is considerable, and at present the cost of rendering a subscriber's account is 10s., accounts being rendered twice yearly. With the introduction of subscriber trunk dialling and periodic metering, all calls will be billed together, with no means of identifying charges for individual calls, and the rendering of accounts at more frequent intervals is therefore desirable. Quarterly accounting is, in fact, being introduced with subscriber trunk dialling. Machine methods of preparing accounts have already been introduced, and developments are in hand for the application of electronic methods to accounting processes.

Progress in the application of electronic techniques has been bound up with the transistor, one of the most important recent developments in its probable effect on the future of telecommunication. To the advantages of electronic methods mentioned above, the transistor adds large savings in power equipment and consumption; for this reason and because of the minute dimensions of the transistor itself, reduced costs follow in accommodation and maintenance. Moreover, these considerations allow of the siting of equipment in places not otherwise accessible.

Provided that transistors are shown to retain their characteristics without variation over long periods, there is a wide field for their application to submarine cable systems. The first transatlantic telephone cable with intermediate repeaters using thermionic valves was laid four years ago, a second has already been laid, and others are being planned. A beginning has also been made on a Commonwealth cable system linking the United Kingdom, Canada and Australia.

The limitation of frequency coverage and therefore of the number of circuits which can be provided by a submarine cable link is imposed by the degree of amplification which can be introduced, and this in turn is limited by the power which can be fed along the cable to supply the intermediate repeaters. It is here that the introduction of transistors holds out the possibility of including many more intermediate repeaters, and therefore greatly increasing the circuit-carrying capacity of the system.

At the present time, on the one hand, we have the potentialities for growth and expansion of the telephone service provided by the technical developments described above, and on the other hand, both demand for telephone service and trunk traffic tinue to rise. The former has evoked a reassessment of telepl growth to an estimate of double the present number of teleph by 1975, and it is likely that even this will be exceeded.

The rise in trunk traffic is even more marked, and v present and projected facilities may seem adequate to cope growth in the near future, new developments are arising tc the resources of the trunk system. The number of chai required by television will undoubtedly continue to grow, the increasing application of computer and data-processing t niques, whereby information is transmitted over telephone cuits to centrally situated computers, may make heavy dema

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CAMBRIDGE ELECTRONICS AND MEASUREMENTS GROUP: CHAIRMAN'S ADDRESS

By H. V. BECK, M.A., B.Sc., Associate Member.

'DESIGN PROCESSES FOR ELECTRONIC EQUIPMENT'

(Abstract *of Address delivered at* Cambridge *6th October*, 1960.)

The word 'design' is capable of many different interpretations. In an engineering context, design processes are concerned with the technical as well as the aesthetic aspects of a final product and form part of a larger evolutionary process which leads from originator to user. In the commercial field, pre-design activities such as market research and the drawing up of contracts culminate in a specification which more or less forms the starting-point of the design processes, while the end can be taken as the point at which the design team is concerned only with the handing over of information to enable the post-design activities, such as production and sales, to go forward. In the research field, the evolution of electronic equipment follows an analogous pattern.

It is an advantage for the designer to enter into the pre-design activities and to write the specification, since this will give him background information that will be useful when borderline design decisions have to be made. He should also keep the whole evolutionary chain under active observation before and during the period of design and be ready to make modifications if significant changes, e.g. in production methods, occur.

A specification is a statement of the essential features of the finished equipment. The electrical characteristics, ambient conditions, geometrical limitations and factors bearing on the post-design phase and past effort and experience are all explicitly or implicitly part of the specification. Thus a large number of parameters are put into a melting pot and the designer's j to synthesize these parameters, giving to each an emp appropriate to the application of the particular equipr Concurrently the limitations of the human mind and the req ments of production make the division of equipment manageable units an essential part of the design process.

It is not feasible to carry out a completely objective synt of a large number of parameters embracing different discipl a subjective process is also necessary. The designer draw his own knowledge and experience, supplemented by a from his colleagues and information from the literature evolve a rough appreciation of the way in which the design work out. He then examines the divisions individually ar a whole in more detail, and by constant reference to the p metric pool and re-examination, a satisfactory design eme

Electronic techniques being very flexible, the designe constantly faced with a large number of alternative paths, he must judge and decide on which to follow rather than a the more objective course of evaluating each alternative ag the initial specification. Thus there is introduced a cre process—judgment and decision on incomplete information

The objective approach, although difficult, should perha given greater emphasis. There is on the whole too much rel placed on the subjective and creative processes, so that elect equipment design is in danger of becoming an art rather th fusion of art and science.

Mr. Beck is in the Department of Physics, University of Cambridge.

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THE FUTURE OF ELECTRICS AND ELECTRONICS IN AIRCRAFT

By The Rt. Hon. The VISCOUNT CALDECOTE, D.S.C., M.A., Associate Member.

(*Lecture delivered at a joint meeting of* The Institution *and the* Royal Aeronautical Society 11*th November*, 1960.)

(1) INTRODUCTION

The importance of electricity—in the widest sense—in aircraft missiles to-day needs no emphasis: Sir George Edwards said ently, 'The electrical system is now a fundamental part of aircraft and without it it cannot fly', and, of course, the dance system of any missile is, in the absence of the human nputer, even more completely dependent on the reliable high reliability and be content with 80–90% reliability of the missile system. In this connection it must be appreciated that, to achieve a reliability of 80% in a missile containing 2000 components, it is necessary for component reliability to be of the order of 1 in 10^4. The reliability related to the numbers of components is shown in Fig. 1. Thus in any modern aircraft or missile reliability of electrical systems is of vital importance, and

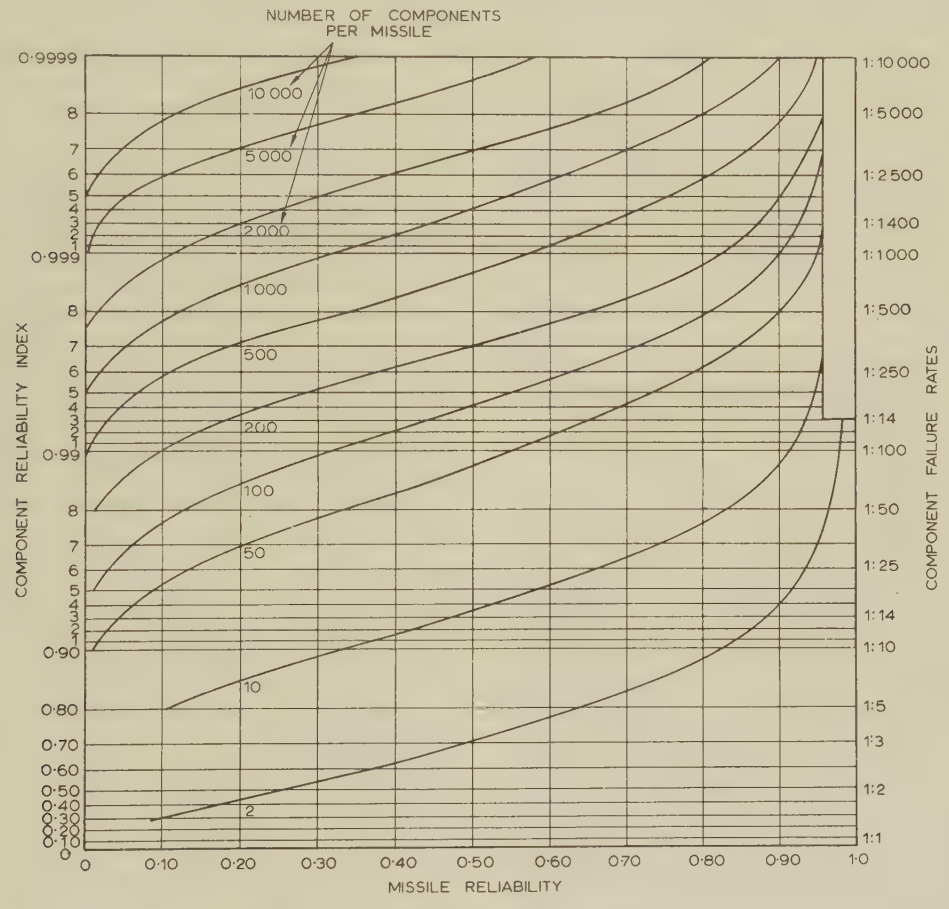


Fig. 1.—Reliability of a missile related to the number of components.

nctioning of thousands of electrical circuits. At present there one important difference between aircraft and missiles. In aircraft the electrical devices must be so designed that at rst no single failure will be catastrophic; either it must be ssible for the human crew to take over or the circuit must be equately duplicated and in the event of failure ensure imme-te transfer of the function to an alternative circuit. In the ssile, however, it is normally impracticable to provide duplica-n which would be of any value, since the failure of any part the guidance and control system will prevent the completion the mission. So we have to depend on using components of in the manned space vehicle of the future this will be even more critical.

There is no need to remind such a distinguished audience as this of the past history of the use of electricity in aircraft, and you will certainly not wish to be taken through a long historical review. Suffice it to say that, from the first installation of simple radiotelegraphy sets and the application of batteries and wind-driven generators for the lighting of aircraft lamps, the use of electrics and electronics in aircraft has reached the stage when large aircraft now have up to 200 kVA of installed power which drives such vital equipment as power controls, fuel pumps, air-conditioning, radio, navigation and flight-control systems.

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It is important to recognize the unity of electrics and electronics. While power generation in aircraft was in the past a purely electrical matter, many electronic techniques are now used in its control, and in both aircraft and missiles it is difficult to draw any hard dividing line in a spectrum of devices from power generation to strictly electronic radio equipment. The central thread in this spectrum is the steadily increasing complexity of all the devices and techniques. It is apparent that, as the performance of aircraft increases and as the air-space fills up, there are more and more functions which the human operator cannot control unaided; in fact, there are many which he cannot control at all, and electronics and electrics must come to his aid. Their function is not merely a convenience to relieve the operator of much of his work, but, increasingly, to do jobs which he cannot do unaided. Having released ourselves from the limitations of a human operator, we now find that we can do much more than assist him—we can virtually disconnect his muscles and his brain and at our convenience distribute them in a way that enables us to obtain information and to perform operations previously inaccessible. This, therefore, is where we stand to-day. Progress demands that both manned aircraft and guided missiles fly higher, farther and faster; in the host of new problems which arise from these demands not the least are those associated with weight and space saving, with safety, reliability and regularity of flight, particularly in adverse environments, and with heat transfer associated with high-speed flight.

(2) POWER GENERATION

It is, I think, convenient to start this general survey with power generation. For manned aircraft four systems have survived from the variety which have been used since the advent of the engine-driven generator less than 30 years ago. These are:

The 28-volt d.c. system.
The 112-volt d.c. system.
The 200-volt 400 c/s constant-frequency system.
The 200-volt variable-frequency system.

Each of these systems has survived because on balance it has advantages over competitive systems: it is an example of survival of the fittest. While prominence is often given to the extremes of operation, we must not lose sight of the need for simpler systems in light aircraft and other short-range passenger-carrying aircraft. For many of these the 28-volt d.c. system will continue to provide an adequate and economical arrangement, especially since developments have now allowed the use of semiconductor regulators with or without magnetic amplifiers, and the advent of the brushless generator has eliminated problems of brush wear. The high currents in such low-voltage d.c. systems, however, will limit their power to about 12 kW per channel, and we must beware lest the attractive developments now possible do not introduce complexities which will detract from the basic economy and reliability of these systems.

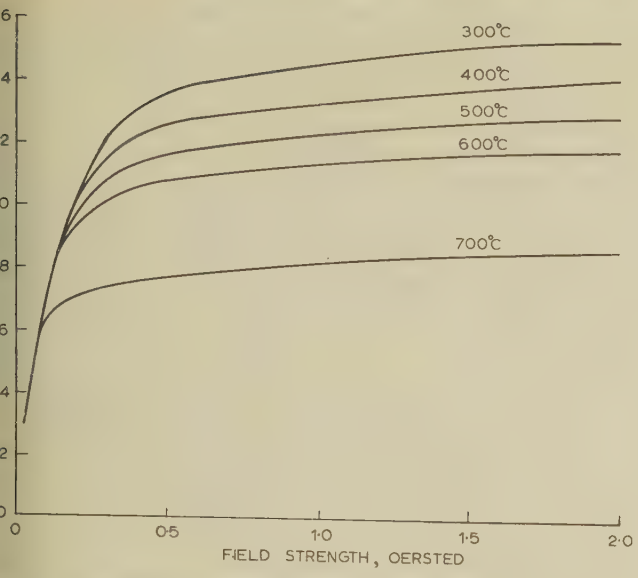
The 112-volt d.c. system is used in some of the earlier V-bombers, but it has several disadvantages inherent in relatively high-voltage d.c. systems, such as radio interference, arcing at contacts and difficult fault clearing. In view of the recent developments in a.c. supplies it is unlikely that this system will survive.

The details of the 200-volt 400 c/s constant-frequency systems will be familiar to all those present; I would, however, like to mention one or two specific problems related to them. At present the first element in such a system is a hydraulic constant-speed drive or an air turbine. From the aircraft designer's point of view it is most efficient to reject the heat generated in hydraulic drives at as high a temperature as possible: at present this is limited by the properties of the oil to about 160° C. Some further improvement in performance coul achieved by an increase in this temperature by the use of e oils. Since these units are now of proved reliability and en any number of channels to be run in parallel, some fu development in this or other directions to increase power/weight ratio might well be worth while. Although next generation of civil aircraft will use this system, c constant-frequency systems using air turbines of one kin another are already in military service and seem likely t used in the future, especially in twin-engine aircraft where ability to operate in parallel is of less value. Quite recen constant-speed drive was demonstrated using an air m differentially to add and subtract speed between engine generator; it is claimed that this drive can also be used a ground air supply as a starter motor or as a motor to the alternators when the engines are stationary. Exte development is, however, needed, since the air available an engine at a sufficient pressure and quantity will be at a 600° C; this may well prove to be as serious a limitatio practice as the temperature range of oil used in hydraulic d

There are many obvious advantages of a constant-frequ system and the ultimate goal is to provide such a sy without the use of complex mechanical constant-speed d with their associated governing and control problems, bu the ability to parallel channels when desirable—the so-c variable-speed constant-frequency (v.s.c.f.) system. The ad of the silicon rectifier and power transistors have made pos the development of static frequency-changers of comparat large power. These can form the basis of a system compr a brushless alternator, driven at engine speed, delivering power at variable frequency to a frequency-changer (contr by a crystal oscillator, tuning fork or tuned circuit) wi constant-frequency output. These v.s.c.f. systems are no the early development stages and seem likely to be lighter more reliable and efficient than any previous system. In e respect they should be most suitable and should be availabl a supersonic airliner, provided that development is act pursued.

The third system in current use is the wild-frequency system. With a brushless alternator this is basically one o simplest systems, but it is suitable only when a large propo of the load is resistive, such as is required for deicing and coo Such a system can also supply power for a.c. motors and c current via rectifiers. Existing electronic equipment norm requires a constant-frequency supply, but the increasing u semiconductors will encourage more use of direct cu This trend, coupled with increasing domestic and deicing l may well bring the wild-frequency system back into fa owing to its basic simplicity.

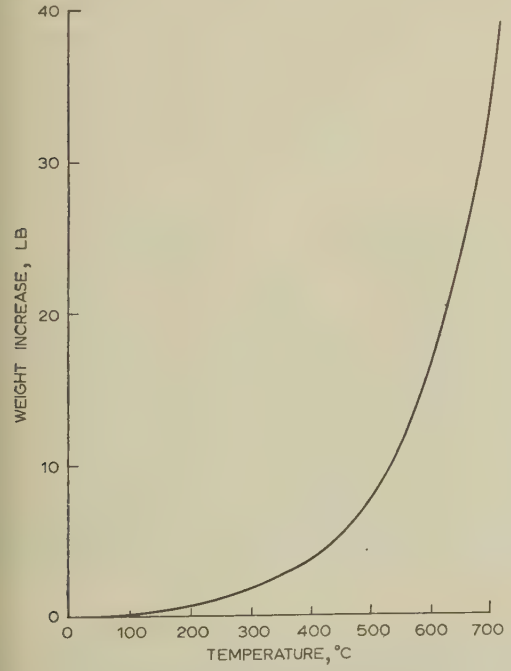
In generators themselves the heat-transfer problem i important factor. The development of gallium-arsenide fiers at least opens the prospect of 300° C operation, and a high temperature for heat rejection has obvious advant there are serious electrical disadvantages. At 300° C w reaching the limit of available insulating materials which the requirements of mechanical strength, stability and s and special precautions must be taken to prevent oxidati the copper under the insulation. The relationship betwee density and field strength is shown for various temperatu Fig. 2. At about 400° C the magnetic properties of iron st deteriorate, and it therefore seems likely that the upper of temperature for the design of generators has nearly reached, and certainly considerable research into the o optimum operating temperature is required. Reliable brus machines, both a.c. and d.c., are now in service, and thes likely to be steadily improved over the next decade withou



ig. 2.—Temperature variation of flux density with field strength.

damental changes. At first sight the solid-rotor alternator ms to have many advantages: the rectifiers are stationary, re is no rotating winding and the operating temperature is h. But experience so far indicates that these machines will heavier than existing ones, and there are many problems in struction, not all of which have been overcome. It is too ly yet to decide to what extent they will replace brushless chines with wound rotors, which are proving reliable and cient, but there are indications in America that the solid-rotor rnator is losing favour; further development and investigation equired. In particular, the increase in core weight with tem- ature (Fig. 3) raises a number of problems.

Electrical power sources in missiles present a largely different blem. The alternatives now being used are direct current m rechargeable or single-shot batteries and alternating current



. 3.—Example of the increase in core weight of a generator with increase in temperature.

from alternators normally driven by hot-gas turbines. Rechargeable batteries or conventional dry batteries are inconvenient, owing to storage problems. However, modern single-shot batteries can be made with a good shelf life, and although they cannot be tested, do provide a reliable source of power and are being used in medium-sized missiles. Improvement in their performance, particularly in low-temperature conditions, would widen their application.

A.C. systems are widely used, even for small missiles with time of flight less than a minute, and it is often convenient to use the prime-mover also to provide power for hydraulic pumps and actuators.

For small and medium-sized missiles the relative advantages of alternators or batteries are fairly evenly balanced and the choice will depend on the particular requirements of each system. For larger missiles with long flight times and greater electrical power requirements, a.c. systems will predominate, while for space vehicles one of the more modern developments mentioned later in the paper will be used.

(3) POWER DISTRIBUTION AND PROTECTION

An important element in the reliability of the electrical system of an aircraft is the distribution circuit, together with its protective arrangements. Owing to their relatively short flight times, these problems are less relevant in missiles at present, but will become vitally important in manned space vehicles of the future. Up to now, wiring and insulation developments have broadly kept pace with increasing aircraft performance, and ambient temperatures encountered in a Mach 2 aircraft can be met using existing insulators based on silicone-elastomer glass-fibre constructions with fluoro-carbon insulators available for particular hot spots. For Mach 3 aircraft with high ambient-temperature conditions the use of wholly mineral-insulated cable and plugs and sockets with ceramic-insulated internals will be required. However, the mating elements of plugs and sockets already represent a temperature barrier, since migration and oxidation place a limit on silvered mating surfaces. These remarks apply equally to crimped connections, terminal tags and the like. Solid nickel surfaces in contact show promise, but the whole subject is one which requires intensive and urgent development, for in both aircraft and missiles plug-and-socket connections are a frequent source of unserviceability. Fortunately the advent of microminiaturization will lead to the combination of more units in one assembly, and therefore to fewer plugs and sockets, and to less wiring exposed to the extremes of environmental conditions.

Protective and other switching devices also require considerable development to improve reliability and reduce weight. Development in two directions is likely: sealed switches, relays and circuit-breakers filled with inert gas are being developed and in the lower-current ranges are already available, although the British equipment industry has lagged behind America in this field. Now similar development is needed on higher-current devices, such as circuit-breakers able to break fault currents of 2 000 amp at 400 c/s. Developments in the atomic-energy field are producing reliable sealed microswitches made of stainless steel and ceramics which will be of value to the aircraft designer for high-temperature applications. The ultimate aim must be to eliminate as many moving contacts as possible, and in the never-ending search for lightness combined with reliability the trend to replace relays and switches by semiconductor devices with no moving parts will continue.

Ultimately, as new circuit and microminiaturization techniques become available, it will be possible to confine a greater proportion of electronic and electrical equipment to areas where

the environment is controlled and some of our present problems will disappear: until this time—which is many years away yet, at least for civil aircraft—development of components able to operate reliably in the ruling ambient condition must be continued.

(4) COMPONENTS AND TECHNIQUES

The progress which has been made over the past ten years in the development of electrical components and techniques has been of the greatest significance in both aircraft and missiles. Much of the stimulus has indeed come from the necessity to compress more and more complex circuits, first into military aircraft, then into guided weapons and now into space vehicles; this is one of the factors which make it essential for this country, not only to continue missile development, but now to take a substantial part in space research. Already civil aircraft and industry generally are benefiting from this progress; high-powered radar, Doppler techniques, semiconductors, computers and microminiaturization are a few examples.

Of all these, microminiaturization probably has the widest application and is likely to have an increasingly important impact in the future. Developments are now proceeding along three distinct lines, namely the flat component, the flat circuit and the solid circuit. Three separate objectives can be distinguished in each of these. First, the production of a high ratio of useful to parasitic mass and volume on the electronic parts; secondly, a minimum of internal hand-made connections; and thirdly an elimination of internal non-productive mass such as sockets, connectors, protective coverings, etc.

The essence of the flat-component technique is to standardize the size and shape of all components. For instance, components can be in the form of circular wafers which are stacked with interleaving insulating material, the circuit being formed from a stack of elements connected on the outside of the wafers. Such units are known as micromodules (Fig. 4).

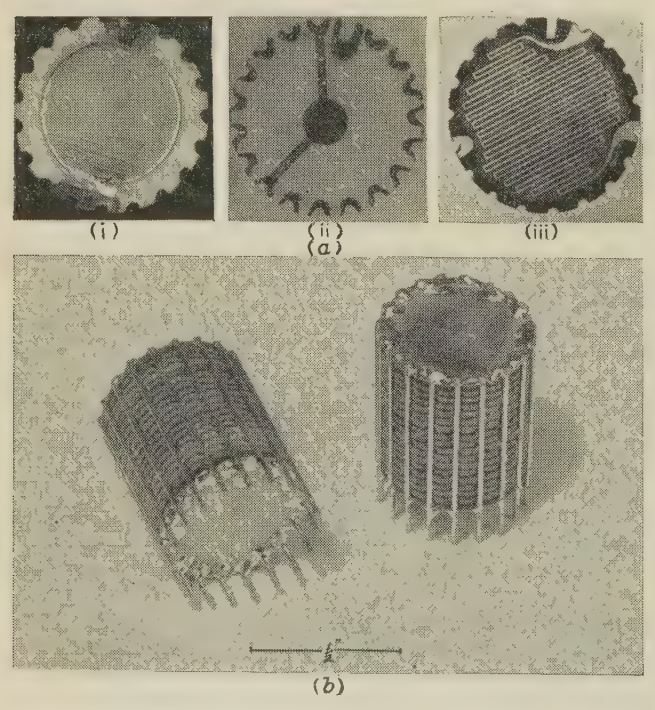


Fig. 4.—Dummy modules.
(*a*) Components of stacks.
(i) 1 620 pF high-permittivity ceramic capacitor.
(ii) 250 pF silicon-dioxide capacitor on glass.
(iii) Resistance pattern on disc.
(*b*) Complete modules (dummy).

The flat-circuit technique involves forming complete circ on the surface of small insulating plates. Active elements s as transistors are inserted like minute aspirin tablets into rece and connected to the remainder of the circuit.

The solid-circuit technique is probably the nearest to ultimate solution of the microminiaturization problem, the being based upon developments taking place in solid-s physics. It has now been found possible to build on a si crystal of semiconductor material multiples of transistors, di and resistors by suitably processing minute areas of the cr surface. Thus, a complete block of circuit is created on semiconductor material itself (Fig. 5). Application of the s

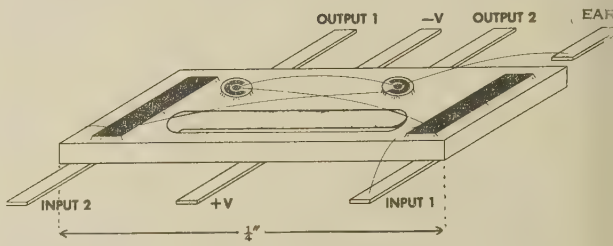


Fig. 5.—Solid-circuit all-silicon multivibrator.

circuit is now being actively studied, but it will be many y before they are an economic proposition in aircraft.

However, the flat-components and flat-circuit techniques well be developed rapidly over the next ten years, altho expert opinions differ on the relative merits of these two meth The former can be used in development where the time f circuit design to manufacture must be kept as short as possi But the flat circuit is a less flexible technique and is more suit for application where changes in circuit techniques are likely, enabling the design to be frozen at an earlier stage, where a manufacturing process can be set up to make relati large quantities. The technical problems associated with t two techniques are also somewhat different: whereas the m difficulty with the flat component is to make the side connecti the problem with the flat circuit is to produce a numbe different components, both reliable and to a close tolerance as to ensure that the completed circuit will be within specificat

Other developments in this field which will become of im tance during the next decade are the parametric amplifier the mesa transistor. The former depends on the principle energy can be injected into an *LCR* resonant circuit by var the capacitance at a suitable frequency. Its advantage is high amplifications can be obtained at very low noise le The mesa transistor takes advantage of diffusion technique obtain a very thin base width for the junction. These transis have an upper frequency limit of about 300 Mc/s at pres although the theoretical limit is much higher. For medium- low-power applications up to, say, 10 watts at 40 Mc/s they h an obvious future in all airborne electronic equipment.

One of the newer semiconductor devices holding promise applications with a weight restriction is the Esaki, or tu diode. The versatility of this device is such that one tu diode will replace the r.f. amplifier, local oscillator, freque changer and first i.f. amplifier in a conventional receiver. (the tuned circuits associated with these stages remain, wi great saving in weight and volume. The fast response of Esaki diode also makes it suitable for use in high-speed comp systems.

The importance of the development of new compone particularly those involving the microminiaturization techni which I have mentioned, cannot be over emphasized. At pres development in this country is not keeping up with that goin

road, and unless this trend is corrected our aircraft and missile signers will be seriously handicapped in the future, or will ve to rely more and more on imported components.

(5) FLIGHT CONTROL, NAVIGATION AND GUIDANCE SYSTEMS

From the earliest days of flying the presentation of adequate ght information to the pilot has been of great importance. itially this was confined to crude readings of indicated air eed, height and fuel state, all from instruments operated ectly without the help of electricity. In those days there was and large plenty of time for the pilot to study carefully formation thus provided and to make his decisions accordingly. ow, increasing speeds have cut down the time available to the ot, particularly in the critical take-off and landing phases; gh fuel consumption and the requirements of air-traffic control ve made accurate navigation essential for both safety and onomic reasons; and the requirements for safe and accurate ing in bad weather have greatly added to the amount of formation required to be given to the pilot by his instruments. This has led to the presentation of information to the pilot a more intelligible form in comprehensive and better inter- ated instrument displays, driven by relatively complex servo stems. So long as they work, these displays are excellent and l the pilot in the simplest terms what he should do. At the ne time, automatic pilots have been developed to provide more d more facilities such as i.l.s. coupling and altitude hold. In these fields developments will continue to provide smaller d lighter equipment, and above all, greater reliability. High- rformance heavy aircraft tend to have smaller safety margins take-off and landing than their predecessors, and this has led a demand for a take-off monitor and improved landing aids, rticularly in bad weather. Developments in both these spheres ll continue, and it is but a short step to fully automatic flight ntrol of all type of aircraft with the crew acting largely as onitors and decision-makers when difficulties arise.

(6) FLIGHT-CONTROL SYSTEMS

It has been shown that, by using multiple systems and com- ring their outputs, a sufficient degree of safety can be achieved the automatic landing of passenger aircraft. Automatic re-out and full automatic landing systems are being developed d excellent progress has been made—quite sufficient to justify provision being made for these systems in the VC10 and H121 aircraft—but it will be a number of years before they e in service.

Controls for vertical take-off and landing (v.t.o.l.) aircraft, ich pose special problems of automatic stabilization, will rtainly require a great deal of thought in the coming years. ecial attention is also required to produce a completely iable blind landing system for all types of v.t.o.l. aircraft to able them to land in any weather on a predetermined spot or an emergency in a ploughed field or similar unprepared ality. Only when this is possible can their full capabilities exploited. Another sphere in which development is urgently eded is that of heading references; at present this is the greatest itation to the accuracy of Doppler systems, and future velopments will be aimed at free or north-seeking gyro systems h long-term drift rates not greater than 0·1° per hour, such are already being used for military purposes.

(7) NAVIGATION SYSTEMS

VOR has been the standard international short-range naviga- nal aid for a number of years and provides track guidance ng a selected radial to (or from) the VOR beacon. Following a decision taken by the I.C.A.O. earlier this year, VOR has been supplemented by DMET, a distance-measuring system, to provide full position information. There is little doubt, however, that the VOR–DMET combination is insufficiently accurate or flexible for high-density traffic areas, and some type of area coverage system will be needed in European terminal areas and the eastern seaboard of the United States. The Decca phase-comparison system could meet this need, but the necessary international agreement has not yet been achieved. No such agreement is needed for Doppler systems, and their very high accuracy over short distances makes possible their use in terminal areas; for this application they will need to be coupled to a pictorial display.

Internationally speaking, there is no agreed long-range navigational aid. Loran and Consol are both agreed as interim systems and, of course, a.d.f. is extensively used. But there is at present no satisfactory ground-based system providing world-wide coverage. Any system requiring a large number of ground stations in different countries leads to severe difficulties in implementation. Work has been going on for a number of years on low-frequency phase-comparison systems, which require relatively few ground stations, but there is still insufficient information on propagation characteristics to be sure that sufficient accuracy and reliability can be achieved. The use of navigation satellites is clearly a promising direction for future development, but for the near future self-contained aids, either Doppler, inertial or—more probably—a mixture of the two, look like becoming the standard systems.

(8) COMMUNICATION

Both in missiles and in manned aircraft communication of information in the widest sense is becoming increasingly important and complex. Signals derived from either outside or inside a system are invariably, to some extent, obscured or debased by background noise. Much development has been, and will undoubtedly continue to be, concentrated on systems which extract the maximum information from signals in the presence of noise, using a given bandwidth and in a given time, and on the development of low-noise signal-amplifying devices.

While it would be dangerous to predict the most likely line of advance, there is no doubt that progress on the first of these—the detection of useful signals in the presence of noise—would be of very wide application. The extraction of weak signals demands correlation techniques which are to some extent analogous to the problem of meeting a stranger at a railway station. If very little is known about the wanted person which can distinguish him from the crowd, the chance of spotting him is small; the more we know about him, the better the chance of distinguishing him. So it is with a signal submerged in a background of noise; if we know something about it we can correlate it with a replica of the expected signal and will thus be able to detect the presence of the expected signal in spite of the surrounding noise, just as we may detect a man with a limp and a white carnation in a crowd. These correlation techniques, however, demand extremely complex circuits containing numbers of matched filters, and the advent of semiconductors and micro-miniaturization will considerably help the designer in this difficult problem. These developments could make an important contribution in both missiles and the automatic transmission of information to and from aircraft.

An extension to this technique was announced in September, 1960, at the Symposium on Information Theory. Three American scientists then described a new technique called 'adaptive waveform recognition' which has demonstrated the ability to detect signals buried in noise and to adapt continually the circuit parameters for optimum reception.

In the past much attention has been paid to the choice of frequency to eliminate unwanted interference and to achieve the most efficient communication channel. This search for new channels will continue at both ends of the frequency spectrum, such as the low-frequency navigation system referred to earlier. For short-range communication civil operators will probably continue to use the v.h.f. band at 50 kc/s channel spacing, while the development of the u.h.f. band will continue for military aircraft. But for longer ranges h.f. single-sideband systems will steadily take over from double-sideband ones during the next decade, giving improved communication and providing more channels per band. Owing to the vagaries of the ionosphere and the severe electrical noise experienced, however, none of the frequency bands currently used for long-range communication is ideal. The best final solution may well be the use of communication satellites allowing the use of v.h.f. and u.h.f. techniques.

(9) DATA HANDLING

Some reference must also be made to the particular problem of data transmission and processing, which are becoming of increasing importance; these will be discussed together, since in many cases it is convenient, not only to transmit data in a form different from that in which it is finally used, but also to combine this data with information derived within the aircraft or missile. The rising density of air traffic and the short flight times of guided missiles both pose the problem of rapid data transmission and the rapid transposing of this data into useable form. In the case of manned aircraft it is becoming quite outside the capability of human air-traffic controllers to assimilate all the information and to make the decisions necessary to ensure the safe and orderly flow of traffic. With increasing speed a similar problem is developing for the aircraft crew, and the automatic transmission, processing and display of information in a convenient form is becoming of great importance; this may well be followed ultimately by the direct control of the aircraft via a data link.

The limits of development in accuracy and miniaturization of analogue computers are likely to be reached quite soon. Noise, drift and limited resolving power all contribute to this. Digital techniques, however, while offering unlimited accuracy, require increasing numbers of digits and consequent increase in circuit complexity. From the airborne standpoint, digital computers offer this accuracy at a very low power level and lend themselves admirably to miniaturization. The real limitation in this case is the time required for any problem to be resolved. In this connection the development of the digital differential analyser is of the highest importance, since it combines the high working speed of analogue methods with the accuracy and ease of miniaturization of digital techniques. A complete digital differential analyser comprises a number of integraters, multipliers and other units using digital techniques, which are connected together in analogue-computer fashion. The trend now in airborne navigation systems is to perform computation by digital differential methods, and this trend will continue for all types of special-purpose computers.

I have concentrated on this need for computers because data processing is probably one of our greatest problems both on the ground and in the air, and it is growing rapidly. Large digital computers are needed on the ground for all the air-traffic-control systems being developed, while in the air, computers are used for flight control and air-data systems, navigation systems and, of course, in military aircraft for the attack and bombing functions. I believe there is a strong case for a central general-purpose computer which could handle all the work required in an aircraft, including such tasks as fuel and flight management.

(10) AIRBORNE RADAR AND COLLISION-WARNING SYSTEMS

As the air space becomes more and more congested a average speeds increase, it is doubtful whether any air-traf control system will be able to ensure complete freedom fr risk of collision. The development of a reliable collisi warning device is urgently required; it must be able to det aircraft which constitute a possible danger in time for relative motions to be analysed, and only those which may ca risk of collision should be brought to the pilot's attenti Ideally, the system should search the surrounding air space a in the light of the information obtained and the 'rule of the roa should indicate clearly to the pilot the action required to be tak to avoid risk of collision. Already visual warning even in go weather is inadequate, and until a reliable warning system least, is available the hazard of collision will increase as the n big jets come into service. The problem is to provide a syst with sufficient coverage without an unacceptable weight a space penalty. Various passive systems dependent on infra-r radiation from engines have been tried, but so far with v limited success.

Airborne radar devices can make some contribution to collision-warning problem, but it is exceedingly difficult arrange for them to cover more than a small arc; however, t can most easily be the forward sector, in association with clo warning equipment, and will give a useful measure of protecti The greatest need is for adequate display of the information a for an analyser to detect the potentially dangerous aircraft.

This is not the place to discuss air-traffic-control problems any detail, but there is one development in this field which m in the future become part of a reliable collision-warning syste Usually, separate radar units are used for surveillance and tracking individual aircraft while they are guided to the fi approach point. The development of track-while-scan ra attempts to combine both these functions into one syste As the radar searches the sky, electronic gating techniques used to examine a selected volume of air space about a particu echo during each scan of the system. Repositioning the track gates at each scan enables the volume around the predic position of the echo to be examined and so enables the relat velocity to be extracted. This repositioning is accomplished computers, and the number of aircraft which can be tracked dependent on the number of computer channels. A sim system might well be developed for airborne collision-warn radar.

Development will continue of higher-powered radar equipm both for aircraft and missiles, combined with even smaller a lighter components. At present the necessity to have la ratios of dish diameter to wavelength to achieve narrow bea of high intensity to some extent limits the value of furt microminiaturization for high-performance radar equipme Any developments to overcome this limitation, such as electric steered beams, would be of great value.

Nevertheless, development will continue of devices capabl producing higher powers at higher frequencies. Probably magnetron has nearly reached the end of its development, the time will soon come when a suitable memorial should erected to the British engineers who developed this small de which played such a vital part in winning the Battle of Atlantic and so saved this country in the last war. It is, h ever, likely to remain the simplest device for the productio high-power beams of microwave frequency, and its use continue for a long time in airborne equipment where freque shifting is not required; where this is a requirement for s reasons as security or electrical beam switching, the amplify klystron will take its place.

(11) MORE ADVANCED PROBLEMS

No survey of this kind would be complete without reference the contribution which electrical techniques are making to e solution of the more advanced problems and ideas which e being actively tackled at this time. On the fringe of the bject under discussion is the whole field of gyroscopes, celerometers and other devices used in inertial-navigation stems. In addition to work on low-friction bearings using nventional principles, investigations have been under way for number of years into radically different designs of gyroscope, ch as gymbal-less gyros in which the rotor is supported with-t mechanical contact by means of magnetic or electric forces. charged body in a static electric field, or a magnetized body a static field, has no stable equilibrium position, and in a ro using this method of support some form of pick-off is cessary to detect both rotor position and attitude of the gyro se relative to the stable platform. Of particular interest in e magnetic suspension type is the cryogenic gyro which makes e of the property of superconduction in certain materials at ry low temperatures, and as such they are repelled by, and us can be supported by, magnetic fields without the need for rvo control. The virtual elimination of frictional torque in agnetic and electrostatic gyros leads to very low drift rates. practical difficulty, however, is that the associated equipment r servo control or supercooling is considerable, while in the ctrostatic gyro it is the very smallness of the supporting force nich leads to mechanical problems.

An even more advanced idea is the particle gyro, which makes e of the angular momentum of rotating particles. The rticle gyro at first sight looks very attractive in that it may be rely electronic. So far, however, this device appears to have ogressed only as far as the initial research stage, and there are ny difficulties to be overcome. There is no doubt that some these developments will make rapid progress during the next n years, so making available gyros with very small drift rates itable for inertia navigation systems in vehicles with long times of flight, such as aircraft and space vehicles. There will also be further progress towards miniaturization of both gyros and accelerometers. With regard to the latter, there are some interesting possibilities using the vibrating-string principle with electrical pick-offs or the piezo-electric effect in crystals, but at present the problems of stable operation in a changing environment have not yet been solved.

An important objective of these developments is to produce an output proportional to acceleration in a form suitable for digital processing. Another approach to this is the pulsed current-balance accelerometer, the success of which is dependent on the production of accurately repeatable current pulses at varying frequency.

A considerable amount of study has already been made of devices for converting heat radiation energy and energy from radioactive isotopes directly to electricity. While at present confined to use in the more exotic forms of space vehicles, within the next decade they are likely to be applied to earth-bound aeronautics. In particular, considerable work is going into thermo-electric generation which could lead ultimately to an aircraft obtaining electric power directly from a hot gas source with very few rotating elements. It seems likely that such power would be fairly low-voltage d.c., and static conversion devices based on transistors and diodes would be required to provide power in an acceptable form. Nuclear auxiliary power units have been under investigation by the American Atomic Energy Commission for a number of years. An American manufacturer has built such a unit, called Snap 1, which develops 125 watts at 28 volts and is based on cerium 144. Owing to the shielding required, the power/weight ratio is very low, being 1 watt/lb. It is reported that a 30 kW unit should be ready within five years. Basically, high energy/weight ratios are available for nuclear fuels—about 1 000 kWh/lb compared with 1–2 kWh/lb for conventional fuels—but owing to practical problems of shielding, temperature of operation and overall efficiency of conversion, it seems unlikely that nuclear power

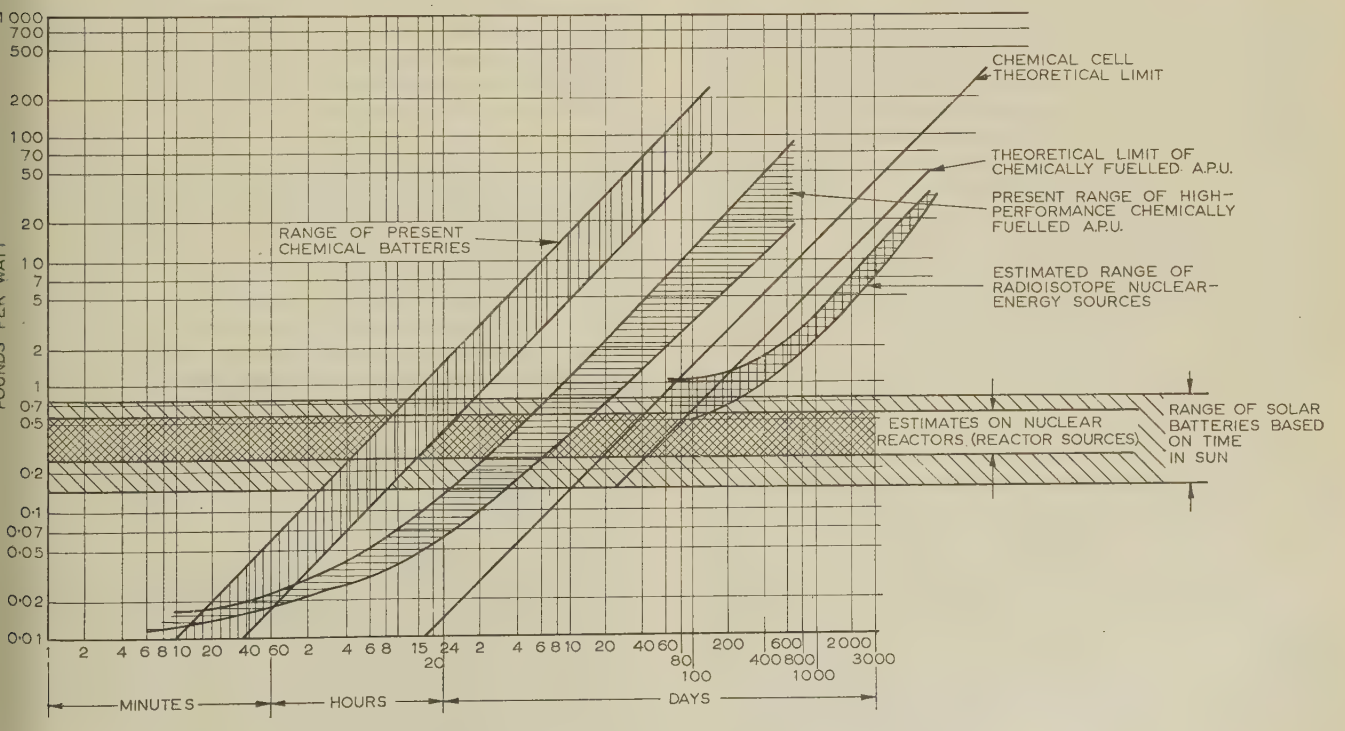


Fig. 6.—Trends in the performance of auxiliary power units.

sources will be used in aircraft for a long time, although they may have an application in space vehicles with very long 'flight times'.

Much work has been done over the past few years on fuel cells, in which electrical energy is produced direct from chemical fuels such as hydrogen and oxygen. The primary advantage of this source of energy is the high operating efficiency obtainable, and a refined fuel cell of this type could have great value in airborne applications, particularly for space vehicles. At present, power/weight ratios are low, but with development these are likely to be increased, making this source of power a practical and useful proposition.

Solar-radiation power is already being used in satellites. With a solar-radiation density of $1 \cdot 25\,kW/m^2$ in space it should be practicable to achieve outputs of 100–150 watts/m^2 using silicon cells of some 15% efficiency, and later 250 watts/m^2 with indium-antimonide cells of 25% efficiency. The performance trends in auxiliary power units are shown in Fig. 6.

There are two other possible methods of airborne power generation which are at present only in the very earliest stages: the first is 'flame generation', which relies on the fact that there is an area on the edge of high-temperature flames which contains a high proportion of charged particles; the second depends on the principles of magneto-hydrodynamics, using the flow of a conducting fluid in a powerful magnetic field.

(12) CONCLUSION

But all this is looking fairly far into the future, and we must not let these dreams occupy our minds to such an extent that we forget the importance of our bread-and-butter business during the next five years. During this time the search for weight- and space-saving, for economy and—above all—for greater reliability of electrical and electronic systems, will continue unabated. Inevitably, complication will increase with greater demands that we make on aeronautical vehicles of types, and the more complex the systems which we install aircraft, missiles and ultimately space vehicles, the greater is need to concentrate on reliability. This is not only for econor reasons, but also because the more functions which are carr out by electrical devices, the more dependent we are upon th for safety of the vehicles as a whole. So I believe that one the most important developments of the next few years, ap from the search for greater reliability itself, is the developm of comprehensive testing arrangements for all the electrical a other devices. Much has already been done in this direction missiles, and the experience should be used in aircraft to prov overall testing arrangements which will both detect a fault a diagnose the cause. This must be achieved without wholes breaking of connections which is in itself such a frequent ca of unserviceability.

As more and more new inventions are produced it is becon clear that, unless there is substantial co-operative effort in western world, it will be very difficult for us to maintain c competitive position against Russia and other countries wh are fast developing the techniques which have been our pres in the past. Even America will not be able to do everyth which is necessary by herself, and I am convinced that more we can get together by technical exchange agreements a such types of co-operation with American and European co panies, particularly in the field of component development, better it will be for all of us.

Finally, I should like to thank all those who have helped in the preparation of this lecture. I hope it will be of so value as a prelude to the discussions between the two societ and that it has provided a few points to stimulate the thoug of those who will take part in these discussions.

DISCUSSION BEFORE A JOINT MEETING OF THE INSTITUTION WITH THE ROYAL AERONAUTICAL SOCIETY, 11TH NOVEMBER, 1960

Mr. C. G. A. Woodford: This lecture gives me great pleasure, particularly because I opened The Institution's first informal discussion on electrics in aircraft during the early part of the war and gave the first of the series of papers on aircraft electrics to The Royal Aeronautical Society in 1945—a paper which has remained fairly prophetic about the development of alternating current for aircraft. Now, with the increasing need for the development and use of data-transmission units and components, I am in charge of the design of data-transmission elements and servo components in the Admiralty for joint Service interests in that respect, and took a party to America for a symposium on the subject in the spring. We were then asked whether we knew what synchros the Russians were using and whether they were turning only to d.c. servos: the author referred to Russia at the end of his lecture.

There are two matters upon which I should like to have the author's comments. The first is the recognition now being given to the vulnerability of the wiring for digital systems. Whereas we might use five wires in a data-transmission synchro system, 15 would be required for digital transmission. We are facing this problem in ships, but conditions must be worse in aircraft in view of the multiplicity of wiring which will ensue from a large amount of digital transmission.

Sir George Edwards: As one who has often complained that designers of electrical equipment are insufficiently aware of the problems inflicted on the aircraft manufacturer when their products prove unreliable or too costly, I think that this lecture is the most forthright contribution to this subject that I have heard. The author referred to the great new developments in technique which are going to be made during the next 10 or 15 years, and he stressed that they must be accompanied by improvement in reliability. This has not always been und stood by those determined to get more amperes out of a squ centimetre. They have succeeded in outdoing their co petitors on performance, but they have not always made same progress in terms of reliability.

Having encouraged this joint approach by our two societ I welcome it very much. It has been very difficult over the ye to get the aircraft and electrical engineer to understand another's problems. Electrics in aircraft and missiles becoming increasingly complex, and we must accept this a continuing trend. There is nothing better than a joint appro to the problem, and I hope that present progress will contin I think that such developments as automatic landing will emp size this. However, co-operation does not want to be confi to a meeting such as this, because only a small part of the airc and electrical designing interests are represented here. Th is a classic opportunity for the two societies, through t branches, to make sure that this missionary work is carried wherever aircraft are made and electrical equipment is desig I hope that the maximum possible impetus will be given to branches to repeat this sort of gathering.

Finally, I should like to ask the author a question which sc may think is unfair. It is obviously going to be very difficul design equipment for increasing temperatures, because nea all of the author's curves look like curves associated with d rises. Can the author estimate the relative difference betw the times it will take to design a set of, say, generating equipm to cope in sustained flight with civil reliability at Mach 2 an second set of equipment with the same reliability for Mach 3

Mr. P. P. Eckersley: I was present when wireless telephony s first transmitted from the ground to the air—the ground ng Brooklands and 'the air' being an aircraft travelling at out 40 m.p.h. Since then I have had a good deal to do with it. During the pioneering days of commercial aviation, a senior cutive remarked to me: 'You have to kill passengers to make ation pay'. Taken literally, the statement is horrifying; by plication it means that some degree of risk must be accepted. other words, is it possible to place too great an emphasis upon ability? It is, of course, a matter of degree. The author has ntly stressed the importance of reliability. May this not be ried too far and so discourage innovation? 'Nothing ture, nothing have.'

Mr. J. E. Serby: My interest is in guided weapons, and I uld assure Mr. Eckersley that we are not afraid of 'killing ssengers'. Our problem is not to make the things so novel d so unreliable that we add hopelessly to the cost of our pro- mme, and that is just as big a problem in many ways as that of ling passengers. I will come back to the question of cost in a ment.

Perhaps I might take up a specific point on which I am in no se an expert. The author mentions flat circuits, but when e tests such devices by putting them into the air, will one not handicapped by the long time and higher cost involved in viding miniaturized circuits when they are reduced to that ent?

The topic of cost is really all-important, and the author knows s as well as any of us. Will not these same devices perhaps d more to the cost of a guided weapon than is justified?

I feel that one of the great things which our colleagues in the ctrical and electronic world could do would be to produce same reliability in a range of articles which cost far less. erything in a guided weapon is expensive. The result is that s extremely difficult to provide the Service departments with many of the missiles as they want, and to a large extent this is cause we are insufficiently cost conscious.

Dr. H. M. Wilson: The author quotes a reliability target of ne 80–90% for the guided missile. When dealing with an craft in automatic operation we must think in terms of the ances of catastrophic failure, which are less than 1 in 10^7 or en 10^8. Moreover, aircraft and equipment manufacturers are nfident that they can attain this level of reliability within a y years.

The advantage which the aircraft designer has over the missile signer probably lies in three main features. First, we can use r vehicle over and over again, so that if something goes wrong s normally possible to find this out and correct it quickly and re can be a steady trend towards improvement. Secondly, gain from size, and in spite of the emphasis which the author ces on miniaturization, it is not always prudent to make mponents as small as possible. Many troubles have been used, for example, by the tendency of designers to use micro- tches in situations where ample space and operating force are ilable for less-sensitive devices.

The third factor which, as the author says, is beginning to y a major part in aircraft guidance and control systems is use of redundancy. In this connection I think that the use highly selective redundancy on the VC10 aircraft is an portant development which is worth following closely, cause it may be uneconomic to achieve the required level of ability by mere multiplication of identical control systems.

Mr. H. Zeffert: The author concentrates rather on the elec- al and electronic reliability in aircraft, but reliable elec- nic equipment would be useless if it was not coupled with chanical reliability. Structural and mechanical design must hand in hand with electrical design to form an integrated whole. Since this is a joint meeting of The Institution of Electrical Engineers and The Royal Aeronautical Society, the thing that we should appeal for is the support of structural and system designers in first helping us to pinpoint the problems which are worthy of solution, and then co-operating in the solution of those problems from both the mechanical and electrical aspects. The VC10 aircraft is an ideal example where electrical and mechanical design have been considered from the inception of the aircraft.

Mr. G. V. E. Thompson: The author considers it essential for this country to continue with costly development and to take part in space research. Will he enlarge on the second part of that statement? Is it sufficient for Britain merely to put satellites into American vehicles, or should Britain—as I think—continue with her own programme of developing launching vehicles as well?

There is one aspect of electrics and aircraft or spacecraft which seems to have been omitted from the lecture, and that is the use of electrics for ion propulsion. In America various types of ion motors are being developed. Will the author comment on this and on whether Britain should undertake the development of these methods of propulsion?

Mr. B. Cape: The author outlines a trend towards aircraft electrical systems becoming more complicated and he has rightly stressed the importance of reliability. In addition to the great technical effort devoted to developing new techniques, it is essential to have well-trained engineers available to maintain equipment when in service. Are the present technical training arrangements going to produce the right type of maintenance engineer? If we cannot be sure of this, changes to the training techniques will have to be made.

Sir George Gardner: I was greatly encouraged to hear about the promise of the new lightweight components which will pack into small spaces, because one of the things that we suffer from in the aeronautical world is 'weight growth', and electrical equipment is not absolved from blame in this connection; all kinds of equipment, as well as the aircraft structure, seem to grow in weight as development proceeds.

I have a great fear that this lightening trend can be interfered with to some extent, and even that it can be destroyed, by the very rapid growth in complexity—not in the sense in which Dr. Wilson was speaking, but in the sense that there is this facility to add together myriads of small components and play the most wonderful tricks with them. Over the years I have been disturbed by the blindness which seems to afflict individuals who find themselves shut away in laboratories with easy access to these myriads of components. They can solve complicated problems, often all too easily; they are probably conscious that if they tried to do the job in another way, say by mechanical means, this would imply making a drawing and getting the components made up in the workshop, and that takes time. The adoption of the line of least resistance in the early stage may therefore lead to a solution which is more complicated than it need be for production and for the customer.

Accordingly I would ask the author to couple with his plea for the pursuit of reliability a plea for additional consciousness of the need to do the job in the simplest possible way. This would make a major contribution to the solution of the problem of reliability.

Mr. A. H. Carter: I suggest that a permanent Anglo-American team be formed, their terms of reference being to co-ordinate the combined resources of both countries for all future defence projects. One of their starting-points would be to examine all future defence requirements of both countries, and then recommend where, and on what, our combined resources can be used. They should also encompass the future technical and financial

resources available. This team could make a completely new approach to the problem of using the resources of the West in the most logical and efficient manner. At the moment there is a terrible waste in having so much duplication of effort.

Mr. M. Settelen: A great many of us have thought for a long time that the future trend in long-range aircraft communication would be toward the single-sideband system; the American Air Force has been using this method for some time. However, looking forward to the future, with the aircraft manufacturers seriously talking about supersonic civil aircraft, there is a very severe and serious problem in using such techniques in aircraft flying at Mach 2 or more. In an aircraft flying at 600 m.p.h. the Doppler shift is 20 c/s at 20 Mc/s. This is apart from any drift in the airborne or ground oscillators. At Mach 2 and above this shift would make it almost impossible to communicate by single-sideband methods unless some means could be devised of automatically correcting it; this would be very difficult, since it would depend on the direction of the aircraft with respect to the ground station at any given moment.

It is my opinion that long-range v.h.f. techniques may supersede s.s.b. ones before the latter are generally adopted in civil aircraft. Considerable research work is being done in this field, particularly by one of the major American international airlines. It is understood that this airline recently decided that all h.f. equipment would be removed from its aircraft within two years and it plans to only use high-power v.h.f. equipment for world-wide international communications. There are obviously problems in the design of such v.h.f. equipment, but these are not insuperable, and I think that long-range v.h.f. systems have a very important future in aircraft communications.

Mr. J. W. McPherson (*communicated*): The author emphasizes the need for weight saving and reliability in all the electrical and electronic equipment in aircraft; he also mentioned four types of supply system, and predicted that one, the medium-voltage d.c. system, would die out. The implication of remarks is that we must make a choice of supply system from of the three others, but are there any good grounds for belie that any of these systems is the ideal or even a near approac the ideal for electronic equipment?

In this context, the ideal supply for any type of utiliza equipment is that system which results in the smallest, ligh and most reliable equipment. Both the d.c. and 400 c/s syst have much to recommend them for the bulk of the electr load. Rotating machinery can readily be made for either and much of the electrical equipment is of this form. electronic equipment, however, usually requires an alterna supply in order that numbers of isolated supplies of diffe voltage levels may be readily derived by the use of transform To achieve minimum weight and volume in the electronic eq ment it can be shown that the supply should be of much hig frequency than 400 c/s, and should preferably have a squ waveform. It should also be free from transient voltage va tions. All existing supplies suffer very seriously from th which are caused by the switching of large power loads which themselves insensitive to such transients.

It may be that the best solution would be to separate airc electrical loads into two categories: all loads which are insc tive to transients could be fed from one supply system, w voltage-sensitive loads such as the majority of electronic eq ment could be provided with an entirely separate sup optimized to provide the minimum size and weight and maximum reliability of the electronic equipment itself. additional weight and complication resulting from the provi of two supply systems could well be more than offset by rec tions in size and weight within the electronic equipment.

On the subject of automatic testing it is worth noting that f automatic testing equipment for the Sea Vixen airborne ra has been designed and is now in service with the Royal Nav

THE AUTHOR'S REPLY TO THE ABOVE DISCUSSION

The Rt. Hon. The Viscount Caldecote (*in reply*): I was very pleased that so much emphasis was placed on reliability in the discussion, since I felt that perhaps this point should have been made more strongly in the lecture.

Mr. Woodford's point relating to the vulnerability of digital circuits is really another case of reliability. The best solution is to minimize the amount of wiring required. In aircraft, where there is less need for digital transmission over long distances, the trend towards miniature components and therefore more compact equipment will go a long way to meeting this point. I cannot agree with Mr. Eckersley that the search for reliability can be carried too far. Reliability is important for its effect on cost, as was clearly pointed out by Mr. Serby in relation to guided weapons, and this is, of course, just as true for aircraft, quite apart from considerations of safety.

I must thank Dr. Wilson for what I take to be his support for my views on the need for redundancy in important control circuits, although I fully accept his view that selective redundancy combined with adequate monitoring may be the best and most economic solution. With Dr. Wilson I would agree that miniaturization for its own sake is undesirable, but both Dr. Wilson's comments and Mr. Serby's fear of increasing cost of missiles due to miniaturization are well answered by Sir George Gardner's comments on the useful part played by miniaturization in preventing 'weight growth'.

Sir George Edwards was most kind in his remarks. He is quite right in drawing attention to the mounting electrical problems associated with increasing speed and resultant temperature rise. I fear that any estimate of development time for a generator operating continuously at Mach 3 would be a guess, but it would in my view take at least twice as long as the deve ment of a machine for use at Mach 2.

I entirely agree with Mr. Zeffert's important point that trical reliability in aircraft is not enough: much unserviceab even in electrical equipment is due to mechanical trouble.

Mr. McPherson's communication on the development dual aircraft electrical system is most interesting. The incr of resistive loads for de-icing, cooking, etc., as pointed out in lecture, seem to make such a system desirable and the us compact and light semiconductor devices may well pave the for its development.

Mr. Settelen's remarks on the effect of Doppler shift single-sideband h.f. communication raised an interesting po At a speed of Mach 2, say 1 500 m.p.h. radially to or from ground station, the shift would be about 45 c/s at 20 Mc/s. would certainly cause an appreciable and annoying loss quality, but would be unlikely to render communica impossible.

Finally, I should like to say a few words in reply to Thompson. Surely the whole question of a British sate launching vehicle is one of cost? This country cannot affor do without a programme of space research and developm but it is sensible that much of this work should be on an in national basis, partly because of the cost involved and pa because it seems to be an excellent field for international operation leading to better understanding between nati Only if attempts to arrange such co-operation fail should decide to 'go it alone'. Similar remarks could also be app to Mr. Thompson's second comment relating to ion propuls which is mainly of value for the propulsion of space vehicles

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PID METHODS FOR ASCERTAINING WHETHER THE ACTIVITY OF A WEAK RADIOACTIVE SAMPLE EXCEEDS A PREDETERMINED LEVEL

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paper was first received 17*th September*, 1959, *and in revised form* 23*rd February*, 1960. *It was published in July*, 1960 *and was read before the* MEASUREMENT AND CONTROL SECTION 25*th October*, 1960.)

SUMMARY

ne paper considers methods of determining whether the mean of occurrence of random events, such as radioactive disintegra-s, is greater or less than some given tolerance rate. The most ient method is one which makes this determination in the shortest and with an acceptably low probability of error. Sequential test edures minimize the sample size by observing the random process some terminating condition is satisfied. The performance of ral sequential tests is investigated, with particular reference to ditions encountered in monitoring radioactive contamination on nands. The most efficient of these tests is shown to be one which sures the difference between the number of events actually observed the number expected at the tolerance rate. Under conditions likely e encountered in routine monitoring for radioactivity on the hands method gives a fivefold saving in time over the method now in use.

LIST OF PRINCIPAL SYMBOLS

λ = True mean occurrence rate of the random events.
λ_c = Tolerance rate.
n = Number of events observed since beginning of test.
t = Time elapsed since beginning of test.

arameters defining conditions under which the test ninates:

n_0 = Predetermined number of events.
t_0 = Predetermined time duration.
K = Predetermined deviation limits above and below the number of events to be expected at the tolerance rate.

(1) INTRODUCTION

Vhen events occur at random the accurate measurement of r mean occurrence rate requires the observation of many nts. A quantitative measurement of the activity of a weak oactive source may therefore take a long time. In some s it is sufficient to make a qualitative test to determine ther the activity is above or below a predetermined level, in principle, this only demands an accurate estimate of the rrence rate when the activity is close to the predetermined l. Sources which are much stronger or much weaker require accurate discrimination and can be tested in a shorter time. equential test procedures take advantage of this possibility test whether the mean occurrence rate of a random process is ve or below a stated rate, in such a way that the sample size ot fixed, but depends on the outcome of the observations. lly the test should terminate as soon as a decision can be n with an allowable probability of error, but practical con-rations make it desirable to choose a less complicated rion which has approximately the required effect.

ne application of sequential test procedures is in the routine surement of radioactivity on the hands. Many pairs of hands must be tested in as short a time as possible at the end of a working shift; most are expected to be uncontaminated or only very lightly contaminated, and even contaminated hands yield quite a low counting rate. An efficient test procedure should allow hands which have very little activity to be tested rapidly, and the same applies to any heavily contaminated hands. It is only those hands whose activity appears to lie close to the tolerance level that need be tested for a longer period in order to give sufficient statistical accuracy.

Most existing hand monitors[1] use the principle illustrated in Figs. 1A and 1B. The counting is effected by accumulating on a capacitor C_1 the charges, Q, produced by each random event, so that the capacitor voltage rises as shown in Fig. 1B. The

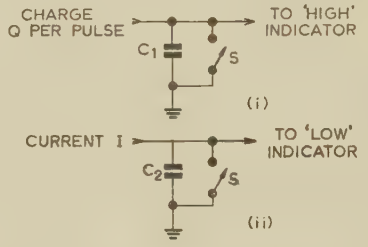


Fig. 1A.—Principle of counting and timing circuits used in most existing hand monitors.
(i) Basic counting circuit.
(ii) Basic timing circuit.

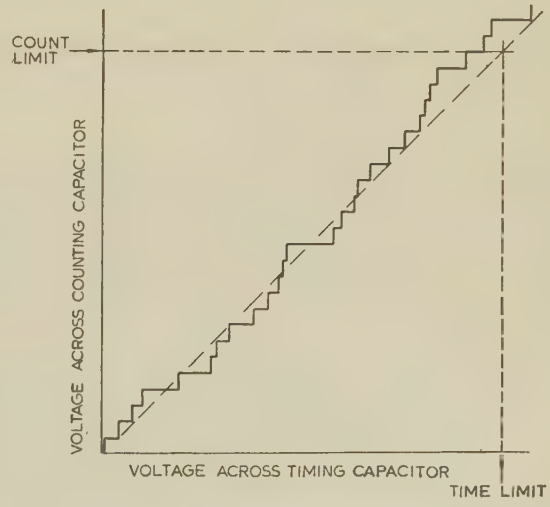


Fig. 1B.—Voltage across counting and timing capacitors when testing the occurrence rate of random events close to the tolerance activity.

authors are in the Electronics Division of the U.K.A.E.A. Atomic Energy rch Establishment.

sampling period is timed by causing a steady current, I, to charge a timing capacitor, C_2. Both capacitors can start to charge when the switch S is opened at the beginning of the period. When either one of these capacitors has acquired enough charge to operate its associated indicator, the appropriate indication is given and the test is terminated. Thus the two circuits impose the count and time limits indicated by the dotted lines in Fig. 1B. This test procedure will therefore be referred to as the rectangular-boundary method. It has the unfortunate characteristic that all tests producing the 'below tolerance' decision (which may be expected to predominate in hand monitoring) require the maximum time allowed by the time limit.

The principles of the faster method[5,9] to be discussed in the paper are shown in Figs. 2A and 2B. Here there is no separate

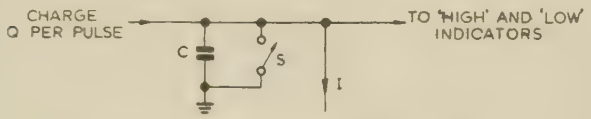


Fig. 2A.—Principle of difference integrator.

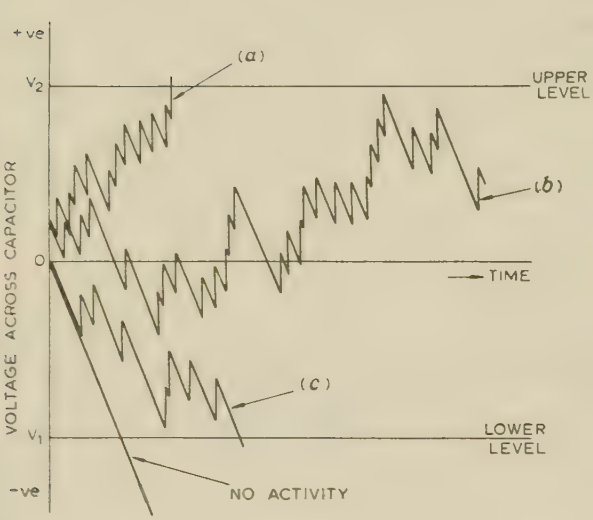


Fig. 2B.—Voltage across capacitor when circuit is used to test occurrence rate of randomly timed pulses.
(*a*) High activity.
(*b*) Activity close to tolerance.
(*c*) Low activity.

timing circuit, but a steady current, I, is caused to flow out of the counting capacitor. This current has a value, $Q\lambda_c$, which would just balance the charge entering the capacitor if the pulse rate were exactly at the tolerance rate, λ_c. Thus the capacitor integrates the difference between the actual pulse rate and the tolerance pulse rate, and the voltage observed across the capacitor at any time after the opening of the switch S represents the difference between the number of pulses to be expected at the tolerance rate and the number which have actually occurred.

If the pulse rate is far above the tolerance level the capacitor charges positively in a series of steps of equal height when S is opened [curve (*a*) of Fig. 2B]. When the voltage reaches $+V_2$ an indicator operates to show that the pulse rate is above the tolerance level.

If the pulse rate is zero, the voltage across the capacitor falls steadily at a rate $Q\lambda_c/C$ (Fig. 2B), and when it reaches a value $-V_1$ another indicator operates to show that the activi below the tolerance level. This indication occurs after a $-K/\lambda_c$ (where K is equal to V_1C/Q, or the ratio of V_1 t instantaneous voltage change caused by a single pulse). It be shown later that K can often be assigned a value of the o of -6, so that, when the activity is very low, a decision is obta in a time equal to that which would be needed to count a six pulses at the tolerance rate. This time is much shorter that for the equivalent rectangular-boundary test.

If the pulse rate is not zero but is below the tolerance l the approach of the capacitor voltage to $-V_1$ is delayed by pulse [curve (*c*)]. The more nearly the pulse rate approache tolerance rate the longer will be the counting period, and accuracy of discrimination will increase accordingly. The s applies to pulse rates just above the tolerance level. If the r counting rate is exactly equal to the tolerance rate the c would continue indefinitely were it not for the random fluc tions in the pulse rate. These terminate the count eventuall causing one or other indicator to operate.

In this test the boundaries are represented by two limits and V_2, and it will therefore be referred to as the pa boundary test.

After discussing the general characteristics of sequential procedures the paper compares the two test procedures w have been outlined above. For the purpose of this compar numerical values appropriate to hand monitoring will be u although the method is applicable to any test where a quali answer is required.

(2) SEQUENTIAL TESTS

A sequential test procedure, leading to the acceptanc rejection of the hypothesis that the mean occurrence rate random process is greater than a given rate, is simply a rul specifying, at each instant of time during the test, which o following three decisions is applicable:

(*a*) To accept the hypothesis and terminate the test.
(*b*) To reject the hypothesis and terminate the test.
(*c*) To continue the test by making further observations.

At the beginning of each test there is not sufficient informa about the random process to justify decision (*a*) or (*b*) decision (*c*) applies. Eventually it is possible to reach dec (*a*) or (*b*) with an acceptable risk of error, and the test is complete.

Since the hypothesis is accepted or rejected on the basis sample of finite size there may be errors of two kinds: eithe hypothesis may be rejected when it is really true, or it ma accepted when it is really false. The relative importance o two kinds of error may not be equal. In the case of monit hands for radioactivity it may well be more serious to ind that heavily contaminated hands are below tolerance tha indicate that lightly contaminated hands are above toler There will also be a range of activity, near the tolerance l where it is immaterial which decision is made, since the mea ment is in any case subject to other uncertainties.

The performance of a test procedure of this type can be s fied by an operating characteristic (o.c.) function which sh how well the test achieves its objective of making co decisions, and an average sample number (a.s.n.) or ave sample time (a.s.t.) function which represents the cost of decisions in terms of either the number of events which mu observed or the time taken to perform the test. Fig. 3 repre the probability of a sequential test terminating with accept of the hypothesis that the actual rate is greater than the toler rate, λ_c, as a function of the true mean occurrence rate, λ. a.s.t. function describes the mean time taken to reach a deci

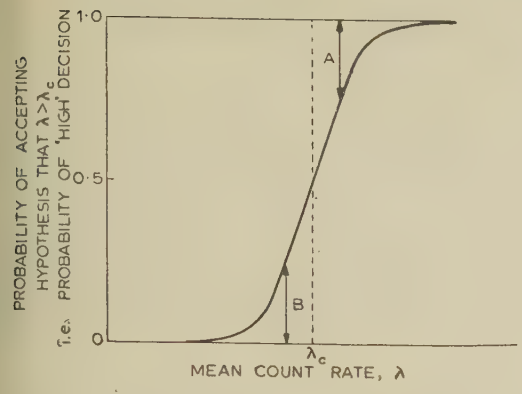


Fig. 3.—The operating characteristic function.
A = probability of error due to rejection of true hypothesis.
B = probability of error due to acceptance of false hypothesis.

n as a function of the true mean occurrence rate of the om process. (For the particular applications which are considered in this paper the average sample time is of importance than the average sample number.)

TEST PROCEDURE WITH RECTANGULAR TIME AND COUNT BOUNDARIES

e rectangular-boundary test procedure in Fig. 1B can be esented as in Fig. 4. The conditions at any time during the

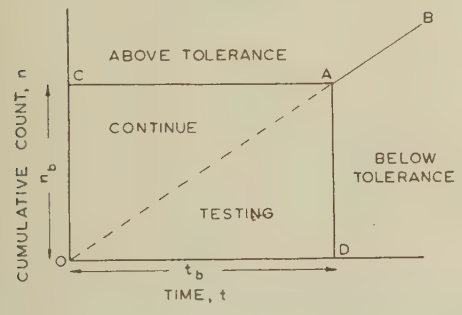


Fig. 4.—Sequential test with rectangular boundaries.

are defined by a point on the diagram corresponding to the opriate point of the 'staircase' voltage of Fig. 1B. The ance rate is represented by the line OAB. The time and t boundaries, AC and AD, divide the diagram into three s. The record of every test starts at the origin O and crosses one of indecision, bounded by OCAD, until it enters the of acceptance, CAB, or the zone of rejection, BAD, and is terminated. The rules for this test procedure can therefore ecified as:

a) If the test has continued for a predetermined time, t_b, reject hypothesis that the observed occurrence rate is greater than the erance rate and terminate the test.
b) If the cumulative total of pulses has reached a predetermined ue, n_b, accept the hypothesis and terminate the test.
c) If neither count nor time has reached its predetermined value, ntinue testing.

erating characteristic and average sample time functions his test procedure have been computed and are shown in 5 for the boundary conditions $t_b = 7$ sec, $n_b = 35$ counts. time and counting rate scales chosen are those appropriate tolerance rate of 5 counts/sec, but these curves and those bsequent Figures can be applied to any tolerance rate by ting the counting rate and time scales in inverse proportion.

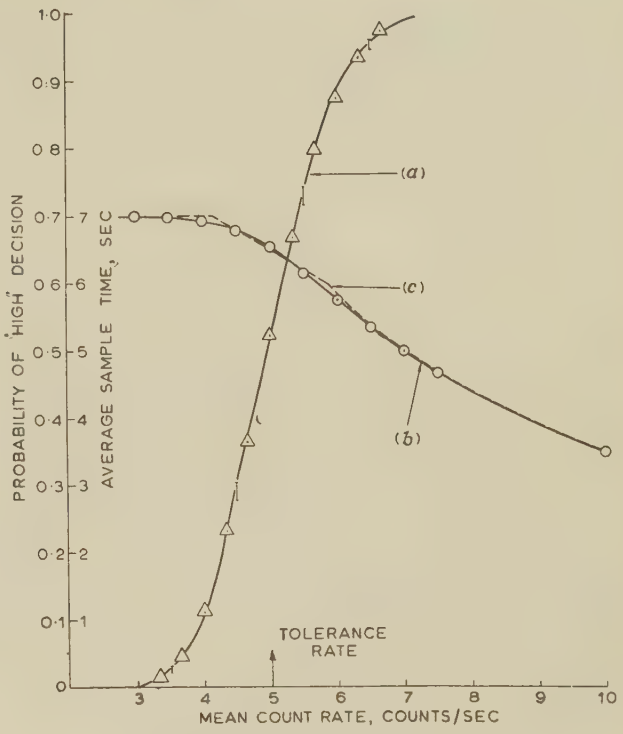


Fig. 5.—Operating characteristic and average sample time for test with rectangular boundaries.
$n_b = 35$ counts, $t_b = 7$ sec.
(*a*) O.C. function.
(*b*) A.S.T. function.
(*c*) Approximation to a.s.t. function used in Section 14.3.
△ Points derived from area under Poisson distribution curve; remainder are from Monte Carlo simulation.

The o.c. function can be derived from the Poisson distribution curve by calculating the area under the curve which lies outside a specified ordinate. If the limits for the test are n_b and t_b and the mean occurrence rate is λ, $[\lambda < (n_b/t_b)]$, the probability of the test giving a false 'high' decision is the probability of observing n_b or more counts in the time t_b. The probability of observing exactly n events in time t is $\varepsilon^{-\lambda t}(\lambda t)^n$, so that $P_\lambda = \varepsilon^{-\lambda t_b} \sum_{i=n_b}^{\infty} (\lambda t_b)^i/i!$. Some points on the o.c. function in Fig. 5 have been computed from this expression.

The performance of this test procedure, and that of others which are described later, were also studied with the aid of a digital computer.[6] A Monte Carlo process, involving the generation of random numbers, was used to simulate the time intervals between successive events and the simulated events were then subjected to the appropriate test procedure. In this particular case the Monte Carlo method provided further points on the o.c. and a.s.t. functions to demonstrate the agreement between the analytical and Monte Carlo methods.

(4) THE WALD SEQUENTIAL TEST, WITH PARALLEL BOUNDARIES

The parallel-boundary test procedure is an application of the sequential probability ratio test of Wald[2] and others. This test is well known[7] as a means for determining, with minimum inspection effort, whether the proportion of defectives in a manufactured product is above or below an acceptance level. In the present application of the test a continuous parameter (time) replaces the original discrete parameter (cumulative number of

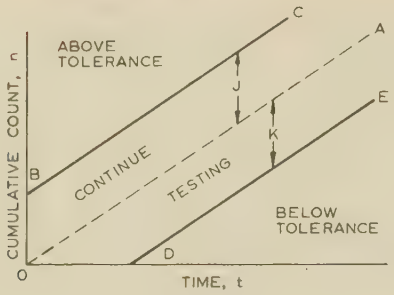


Fig. 6.—Wald sequential test.

items inspected). The Wald test may be represented by a diagram (Fig. 6) in which the co-ordinates are cumulative count and time. The zone of indecision lies between BC and DE, which are parallel to OA and are separated from it by distances corresponding to J and K events respectively. (The principle of the test is more easily demonstrated if the slope of OA, and hence that of the boundary lines BC and DE, is equal to the tolerance rate, λ_c. This convention will be assumed throughout the paper, but Section 14.1 introduces a more general interpretation of the slope of the boundaries.) It can be seen that this is a restatement of the operation of the electrical circuit (Fig. 2B) and is also closely related to the test procedure of Fig. 4.

The rules for the parallel-boundary test procedure may be defined as:

(*a*) Accept the hypothesis that $\lambda \geqslant \lambda_c$ if $(n - \lambda_c t) \geqslant J$.
(*b*) Reject the hypothesis that $\lambda \geqslant \lambda_c$ if $(n - \lambda_c t) \leqslant K$.
(*c*) Continue testing if $J > (n - \lambda_c t) > K$.

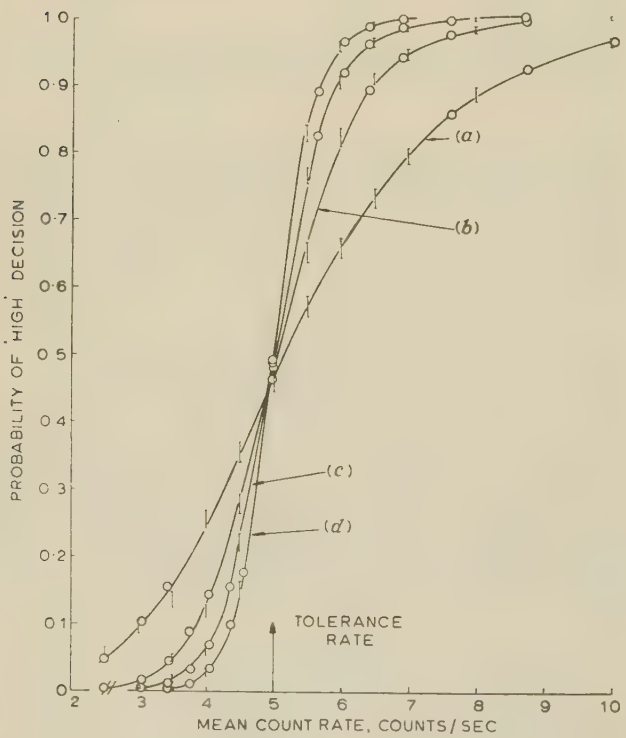


Fig. 7.—Operating characteristic for parallel-boundary test.

Boundary values (*a*) $J = -K = 2$
(*b*) $J = -K = 4$
(*c*) $J = -K = 6$
(*d*) $J = -K = 8$

○ Derived from Kiefer and Wolfowitz.[3]
⌶ Derived from Monte Carlo simulation.

The test is thus closely related to the sequential proba ratio test in which similar rules accept the hypotheses:

(i) $\lambda = \lambda_2$, where $\lambda_2 > \lambda_c$ and
(ii) $\lambda = \lambda_1$, where $\lambda_1 < \lambda_c$.

Tables of the o.c. and a.s.n. functions for this latter tes given by Kiefer and Wolfowitz[3] and can be applied to the pr problem (see Section 14.1). These Tables and the assoc analytical treatment of the problem provide a quick and venient means of designing the basic test procedure for applications. Investigation of modified forms of the test cedure, which will be described later, is carried out more veniently by applying the test procedure to simulated ran events generated by a digital Monte Carlo process. In ord

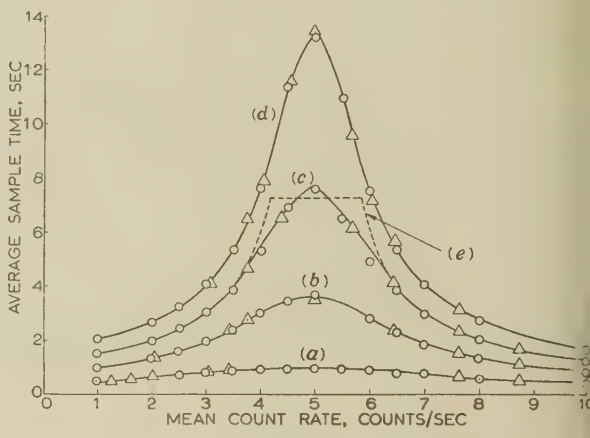


Fig. 8.—Average sample time function for parallel-boundary t

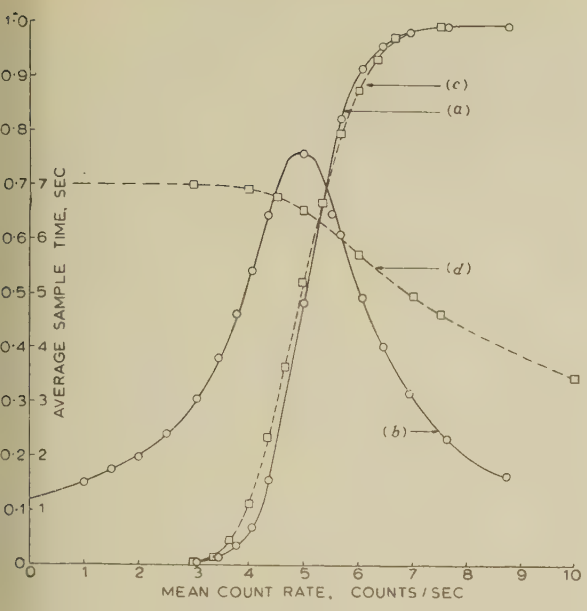
(*a*) $J = -K = 2$
(*b*) $J = -K = 4$
(*c*) $J = -K = 6$
(*d*) $J = -K = 8$
(*e*) Approximation to a.s.t. used in Section 14.2.
△ Derived from Kiefer and Wolfowitz.[3]
○ Derived from Monte Carlo simulation.

establish the validity of the computer programme for simulation it was also applied to the basic case, and Figs. 7 8 show close agreement between the o.c. and a.s.t. tions obtained from Reference 3 and from the Monte C computation.

(5) COMPARISON BETWEEN SEQUENTIAL TESTS WI RECTANGULAR AND WITH PARALLEL BOUNDARI

It is convenient to compare the a.s.t. function for the angular-boundary test in current use (Fig. 5) with that f parallel-boundary test of similar discriminating power, i.e. a similar o.c. function. Although it is not possible to ch boundary values which give identical o.c. functions for the test procedures, the conditions illustrated in Fig. 9 repr substantially similar discriminating power for most prac purposes. The boundary values are so chosen that both give about the same probability of error at true mean c rates 40% above and below the tolerance rate. At count still further removed from the tolerance rate, the rectang boundary test has the smaller probability of error, but here probabilities are well below 1%. For example, when $\lambda =$ rectangular-boundary test has 0·002 probability of error, pared with 0·003 for the parallel-boundary test. Similarly $\lambda = 2{\cdot}5$ the respective probabilities of error are 0·000 07 0·000 36.

The importance of errors is some function of both their nitude and their probability, and this function depends o



).—Operating characteristic and average sample time functions or parallel- and rectangular-boundary tests of similar discriminating power.

○---○ Parallel boundaries; $J = -K = 6$
(*a*) O.C. function.
(*b*) A.S.T. function.

□---□ Rectangular boundaries; $n_b = 35$, $t_b = 7$
(*c*) O.C. function.
(*d*) A.S.T. function.

ose for which the test is intended and on the other possible es of error. The magnitude of an error increases linearly the amount by which λ differs from λ_c, but when this ence is considerable the probability of error decreases much rapidly than the magnitude of the error increases, and it is ned here that the net effect is to decrease the importance of errors. The acceptability of any such test depends upon cceptability of this assumption.

mparison of the a.s.t. functions in Fig. 9 shows that, for ikely distribution of occurrence rates among the samples n are tested, the parallel-boundary test has the shorter ge sampling time.

ctions 14.2 and 14.3 give approximate methods for caling the average sampling time for a rectangular distribution tivities between samples. If such a distribution extends zero to twice the tolerance rate the average sampling time he rectangular-boundary test in Fig. 9 is 5·9 sec and for the lel-boundary test of comparable discriminating power it is sec.

the case of hand monitoring (where most hands are expected uncontaminated) a more realistic distribution would be uncontaminated and 10% uniformly distributed between and twice tolerance. With this distribution, the rectangulardary test in Fig. 9 requires an average time of 6·89 sec per n, compared with an average time of only 1·41 sec for the lel-boundary test. Thus, in this typical situation, the lel-boundary test is on average about five times as fast as ctangular-boundary test in current use.

ce the parallel-boundary test has no inherent time limit, individual tests in which the activity is close to the tolerance can last longer than the corresponding rectangulardary test. In order to ascertain the importance of this, istribution of sample times was investigated in detail for a lel-boundary test with $J = -K = 5$ by using the computer nerate sample times for a series of approximately 1 000 tests.

The worst case was chosen, in which the true mean rate in every test was equal to the tolerance rate. The frequency distribution diagram shown in Fig. 10 was prepared from these sample times. The average sample time at the tolerance rate was 5·3 sec. Nearly 90% of the tests were completed before 11 sec and 96%

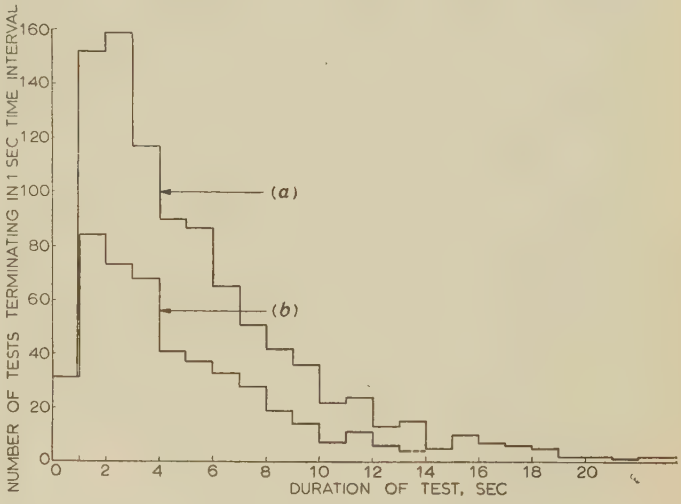


Fig. 10.—Distribution of sample times for parallel-boundary test at tolerance count rate.

$J = -K = 5$; $\lambda = \lambda_c = 5$ counts/sec.
(*a*) Tests terminating at either boundary.
(*b*) Tests terminating at upper boundary only.

before 16 sec. The average sample time of the same test procedure at low occurrence rates is only 1 sec.

The investigation of the distribution of sample times was extended over a range of occurrence rates and to several values of the boundaries J and K by observing the fraction of simulated tests which had a sample time greater than specified limits. Fig. 11 relates to the boundaries ($J = -K = 6$) which provide discrimination similar to that of the 'standard' rectangular

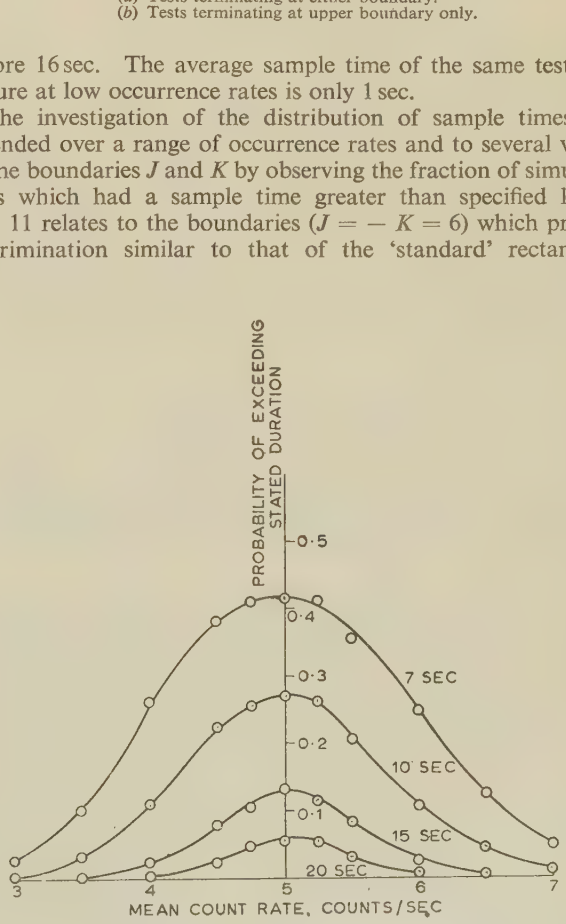


Fig. 11.—Parallel-boundary test ($J = -K = 6$): probability of exceeding stated duration.

boundary test ($n_b = 35$, $t_b = 7$), and shows that the tests which extend beyond the times 7, 10, 15 and 20 sec are quite well confined to a small range of occurrence rates near to the tolerance rate. The maximum probability of exceeding these times occurs when the random occurrence rate equals the tolerance rate. Fig. 12 shows how this maximum probability

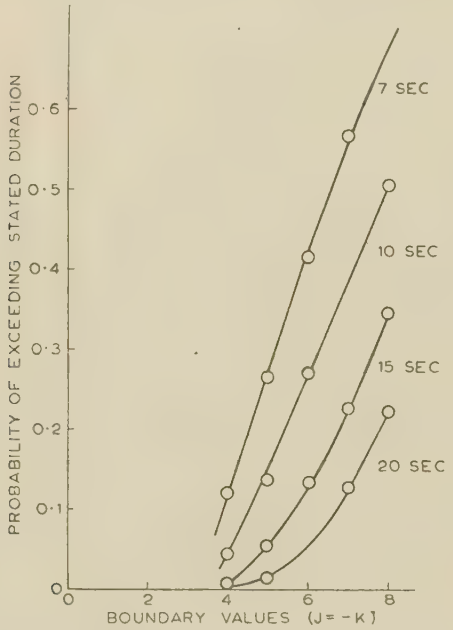


Fig. 12.—Parallel boundary test: probability of exceeding stated duration at a tolerance counting rate of 5 counts/sec.

of exceeding stated sample times is related to the boundary values used.

Although long sample times are so rare that they add little to the average sample time, their occasional occurrence may prove unpopular in personnel monitoring. Moreover, the results they finally provide are of relatively small significance, since their long duration will in itself be a strong indication that the rate being observed is close to the tolerance rate.

(6) MODIFIED SEQUENTIAL TEST PROCEDURES WITH RESTRICTED DURATION

This Section examines several methods by which a time limit may be applied, and shows that the desired effect of restricting the sample time without significant change of the operating characteristic can be achieved by simple modifications to the test procedure.

One method of limiting the duration of the test has been suggested by Wald.[2] If a test has not produced a decision within a fixed time the trend of the test is estimated and used to determine whether the 'above' or 'below' decision is appropriate. Tests which would otherwise be of long duration are thus terminated at the time limit and give a decision of poorer accuracy.

The trend can be ascertained by testing, at the time limit, whether the count of observed events is greater or less than the number expected at tolerance rate, i.e. by testing whether the time/count plot of Fig. 6 lies above or below the tolerance rate line. From Fig. 11 it will be seen that, even if the sampling period is terminated as early as 10 sec (which is only 30% more than the a.s.t. at the tolerance rate), the probability of the time limit being reached is quite small when the samples have a distribution in which low activities predominate, so although less reliable decisions are made when the time limit is rea the effect on the operating characteristic can be expected t slight. A study on the computer showed, in fact, that w 10 sec limit the operating characteristic was almost identical that of the unrestricted test having the same boundaries, shown by the middle curve of Fig. 13. The a.s.t. nea tolerance rate is considerably reduced, as shown in Fig. 14,

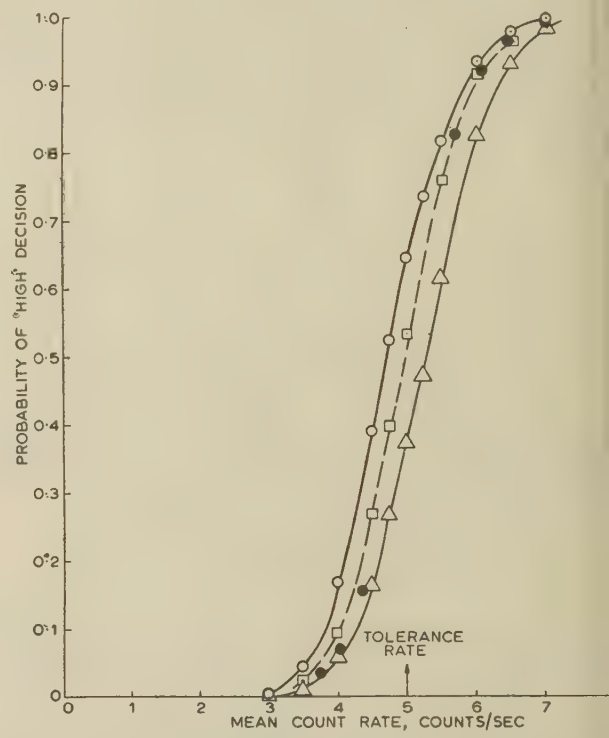


Fig. 13.—Effect on operating characteristic of various methods applying a time limit.

Parallel boundary test ($J = -K = 6$), time limit = 10 sec.
○ All time-terminated tests give 'high' decision.
□ Trend estimation applied to time-terminated tests.
△ All time-terminated tests give 'low' decision.
● Points on operating characteristic of unlimited test of Fig. 9.

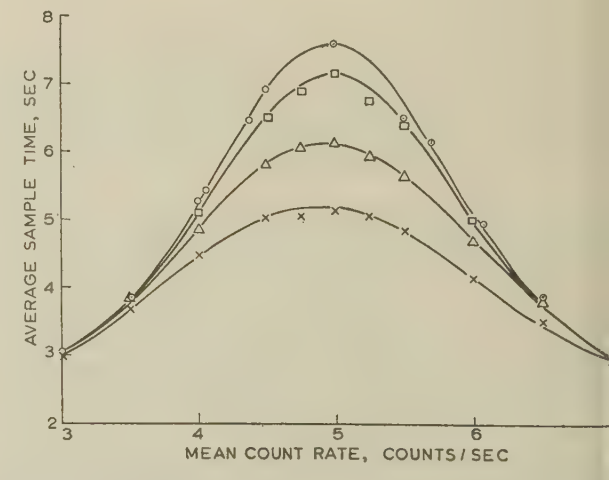


Fig. 14.—Parallel-boundary test ($J = -K = 6$) with time lim a.s.t. function.

× Time limit = 7 sec.
△ Time limit = 10 sec.
□ Time limit = 15 sec.
○ No time limit.

ɔurse, there are no long sampling periods. (A reduction of ime limit to 7 sec was investigated briefly, and this appeared ause considerable deterioration of the operating charac- tic.) A truncated test with an estimate of the trend at the limit therefore appears to be a satisfactory solution to the ɔlem. For most practical purposes the design data for the c test[3] will be adequate for this modified form of the test, in ch the boundary separations J and K have the normal value n $t < t_b$ and collapse to $J = K = 0$ when $t = t_b$. It should be difficult to incorporate this modification in a practical em.

n alternative procedure is to use boundaries which converge tinuously with time. A test with linearly converging ndaries was studied rather extensively, since it was thought ɔffer a compromise between the rectangular- and parallel- ndary tests, providing a rapid decision for samples of low vity and an inherent time limit for samples near the tolerance . The initial separation and rate of convergence of the ndaries were chosen so that both the a.s.t. at low counting s and the operating characteristic were similar to those of the llel-boundary test. The reduction in maximum sampling by this method was found to be less effective than that ieved by the trend-estimation method. Of the two methods, d estimation is also preferred on the grounds that it is ɔably simpler to achieve and does not require special design .

further method of using the parallel-boundary test with fixed limits would be to give a 'near tolerance' decision when a is time-terminated, instead of estimating the trend. With a system it should be practicable to impose a time limit shorter than those in Fig. 14. This version of the parallel- ndary test may be of interest where it is thought desirable to ect people whose contamination level over a period of a week, example, tends to be undesirably high although not actually eeding tolerance. Some users of conventional hand monitors ɔrd the indicated activity as a means of revealing these ginal cases. It should be possible to choose the boundaries time limit so that marginal cases receive a significant number near tolerance' decisions during a week.

still simpler method of time limitation in parallel-boundary s is to treat tests which reach the time limit as if they had all hed one count limit. This method has not been fully lied. Since the trend is ignored at the end of the counting od, some available information is rejected and the operating racteristic must be somewhat inferior to that of the trend- nating method described earlier. There is also a shift in operating characteristic, as indicated by the two outer curves ig. 13. The shift can be eliminated by making the rate at ch charge is withdrawn from the counting condenser . 2A) differ slightly from the tolerance rate, and the resultant rating characteristic is then only slightly inferior to that ined by trend estimation at the time limit.

(7) SEQUENTIAL TESTS IN THE PRESENCE OF A BACKGROUND COUNTING RATE

has been shown that the parallel-boundary sequential test ster than the rectangular-boundary test of a similar dis- inating power. The condition of a low counting rate erimposed on a relatively large background rate will now be sidered briefly. This situation seldom arises in α-particle nting, but is often important in β-particle counting, since rticle detectors are sensitive to γ-radiation entering the rument from outside. In hand monitoring the tolerance nting rate for β-active contamination is considerably higher that for α-activity, but the value of 5 counts/sec will again be used here in order to facilitate comparison with results already obtained and demonstrate that the advantages of the parallel-boundary tests are maintained in the presence of a background counting rate.

Consider the case where the mean background counting rate is known, constant, and is equal to 20 counts/sec. The unknown counting rate is superimposed on this so that the total counting rate when the activity is at the tolerance level is now 25 counts/sec. To obtain the same discriminating power relative to the unknown counting rate, the discriminating power relative to the total counting rate must now be five times as great. To obtain a fivefold increase in statistical accuracy would involve counting 25 times as many pulses. Since the tolerance counting rate is increased fivefold the actual counting time for pulse rates near tolerance would be increased fivefold. Thus, in the rectangular-boundary test the count limit must be increased 25 times (to 875 counts) and the time limit, five times (to 35 sec). Section 14.4 shows that the discriminating power of a parallel-boundary test is directly proportional to the boundary values chosen. Thus, to obtain a fivefold increase in discriminating power the boundary values are increased from ± 6 to ± 30. Fig. 15 shows the o.c. functions for the parallel- and

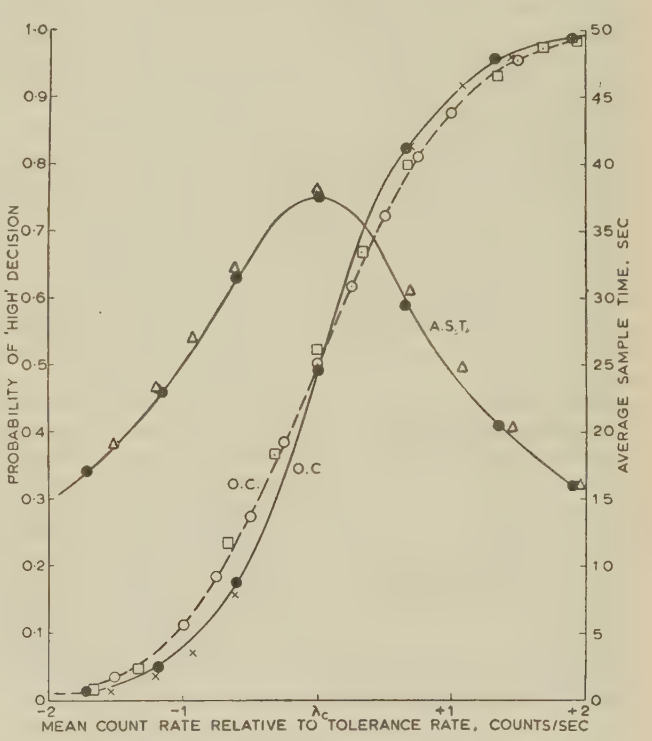


Fig. 15.—Operating characteristics and average sample times for $\lambda_c = 25$ counts/sec.

●—● Parallel-boundary test ($J = -K = 30$), $\lambda_c = 25$.
○---○ Rectangular-boundary test ($n_b = 875$, $t_b = 35$), $\lambda_c = 25$.
× Points on o.c. of parallel-boundary test ($J = -K = 6$), $\lambda_c = 5$, for comparison.
△ Points on a.s.t. of parallel-boundary test ($J = -K = 6$), $\lambda_c = 5$, with time scale multiplied by 5.
□ Points on o.c. of rectangular-boundary test ($n_b = 35$, $t_b = 5$), $\lambda_c = 5$, for comparison.

rectangular-boundary tests which result from choosing these boundary values. Points from the corresponding o.c. functions of Figs. 5 and 9 are also plotted for comparison. It will be seen that increasing the boundary values has caused only minor changes in the shapes of the operating characteristics and that the relationship between parallel- and rectangular-boundary tests is also little changed, both giving about the same proba-

bilities of error at counting rates of 2 counts/sec above and below the tolerance rate, as in the case of Fig. 9. Fig. 15 also shows the a.s.t. curve for the parallel-boundary test and compares this with points taken from the corresponding curve of Fig. 9, multiplied fivefold in time scale. It will be seen that, in this case at least, the average sample time is almost directly proportional to the boundary values. The average sample time for a rectangular distribution of activities between background (20 counts/sec) and background plus 10 counts/sec can be calculated as shown in Section 14.2 and is equal to 16·7 sec. This is just five times the figure of 3·35 sec without background already calculated. From Section 14.3, the corresponding figures for the rectangular-boundary test are 33·4 sec as compared with 5·9 sec without background.

Taking the more realistic case in which 90% of the samples are uncontaminated, we have for the parallel-boundary test:

With background, 7·08 sec
Without background, 1·41 sec

and for the corresponding rectangular-boundary test:

With background, 34·8 sec
Without background, 6·89 sec

Thus, the fivefold advantage possessed by the parallel-boundary test when there is no background is maintained when background is present. The average time saved by using the parallel-boundary test is 5·5 sec per sample without background and 27·7 sec per sample when background is present.

Section 14.4 shows that in a parallel-boundary test having a given discriminating power, the difference between the upper and lower boundaries is proportional to the total of background and mean tolerance counting rate. This suggests that a convenient way of compensating for background counting rate is to use the measured value of this rate to control the amount of charge per pulse fed into the integrating capacitor in Fig. 2A, since this is equivalent to adjusting both the sum of background and tolerance rates and the separation between the boundaries.

A more detailed consideration of the application of both tests in the presence of a background counting rate must take into account the fact that the background rate may vary, so that it cannot be compensated by manually setting the total tolerance rate at some new value. It will therefore be necessary to monitor the background rate, either by registering the counting rate between tests or by using a separate radiation detector, and to use this information to correct the total mean tolerance rate. In both cases there will now be some error in estimating background, due both to statistical uncertainties and to failure to follow any rapid changes in background rate. This will worsen the accuracy of both rectangular- and parallel-boundary tests.

(8) PRACTICAL ASPECTS OF SEQUENTIAL TESTS

From Figs. 1 and 2 it will be seen that the procedures discussed use integrators connected to trip devices which operate when the contents of an associated integrator reach a predetermined level. The integrators may take a variety of other forms such as digital pulse-counting circuits, electro-mechanical devices or even purely mechanical ones. If capacitors are used as integrators they are likely to be in the feedback circuit of a d.c. amplifier, the charges or current being fed in or out at the virtual-earth point. The trip devices may again take many forms, such as bistable electronic circuits, relays, mechanical contacts or photocells operated by the movement of a meter needle.

The rectangular-boundary circuit in Fig. 1A requires two integrators and two trip devices. The difference integrat Fig. 2A requires only one integrator and two trip devices so can be said to be even simpler.

The situation becomes more complex if time limitatio applied to the parallel-boundary system. The timer calls second integrator and a third trip device, but both can be crude since large variations in the timing will have negli effect on accuracy. In general, it will be necessary to arr interlocks between the trip devices so that only one oper Trend estimation may present further problems, though the at least one simple electronic circuit which achieves this. C it will be sufficient for the subject to estimate the trend a end of the sampling time by observing the integrator vo displayed on a meter.

In practical hand monitors there may be as many as counting circuits operating simultaneously, since each hand be monitored by two separate circuits, one counting α-part and one counting β-particles. Counting should continue all four counting circuits reach a decision if there is n limitation. This calls for four integrators and four pa trips interconnected so that counting ends when one trip o pair has operated. Time limitation applied in the obviou would require a further integrator and trip, but the infre occurrence of 'above tolerance' decisions allows the time-li system to be simplified so that the number of trips is re from nine to five. The arrangement is indicated in Fig.

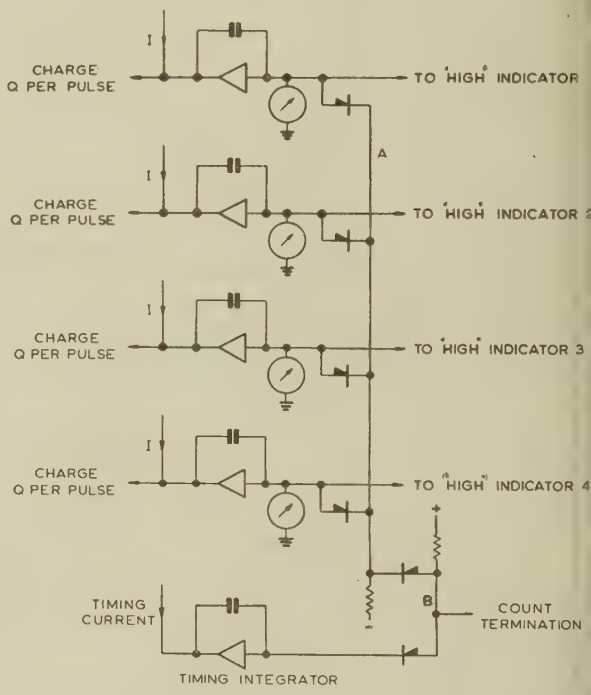


Fig. 16.—Assembly of four difference-integrators with time limi

Each integrator consists of a d.c. amplifier with the integra capacitor in the feedback path. The four difference integra are arranged so that at each pulse the output voltage steps i positive direction, and between pulses it drifts in the neg direction, as in Fig. 2B. The output of the timing integ drifts continuously negative during the sampling period.

The four difference integrators are connected via diodes busbar, A, which therefore takes up the potential of the positive difference integrator. (In computer parlance the di act as an 'and' gate for negative-going signals.) The co

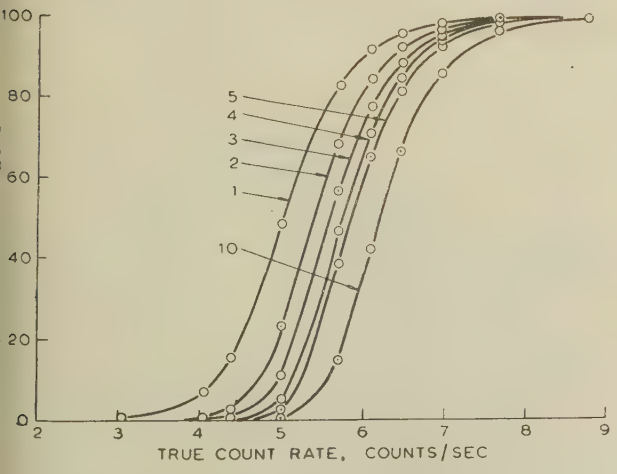
inating trip is connected through two further diodes, which is an 'or' gate whose output follows the potential of the ar or the time integrator, whichever is the more negative. effect is that, if the counting rate in all channels is well below ance, the sampling period is terminated when all the dif- ce integrators have reached the lower limit. The a.s.t. ion is therefore that of the most active channel. If any nel fails to reach the low limit, the sampling period is inated by the time integrator. (It could be arranged that ampling period is terminated if all difference integrators ate their 'high' trips within the time limit, but it is to be d that this situation would be rare, and the total time saved d therefore be small.) Channels which reach no decision n the time limit can be dealt with by any of the methods ssed in Section 6. The simplest in this context is the method ioned at the end of that Section—to consider that all nels have reached a 'low' decision if they have not reached gh' decision within the time limit.

e system in Fig. 16 requires the same number of integrators trips as the rectangular-boundary hand monitors in current ut has, in addition, the arrangement of diodes which allows ampling period to be terminated, not only by the timer, but by the output of the difference integrators.

REPETITIVE APPLICATION OF A SEQUENTIAL TEST

is Section examines the performance of a monitoring me employing a sequential test procedure when the person has received the 'above tolerance' warning presents himself a further test without having reduced his contamination. situation can arise because washing has not reduced the amination, or from a deliberate attempt to repeat the test a 'safe' indication is obtained, or from a genuine distrust e very rapid warning produced by a high counting rate. ny case it is desirable that several repetitions of the test ld not seriously increase the risk that gross contamination receive a 'safe' indication.

the probability of receiving the 'above tolerance' indication when the true count rate is λ, and those rejected from each enter the next test with unchanged activity, the proportion ted after the nth test will be P_λ^n. Values of P_λ can be ined from the o.c. function of the test procedure and used to a series of curves, as in Fig. 17, showing the fraction which have been rejected after a given number of trials. The procedure for this example has parallel boundaries



17.—Performance of a sequential test procedure when rejected samples are tested again by the same procedure.

($J = -K = 6$) and tolerance count rate $\lambda_c = 5$ counts/sec. When the rejects have been tested ten times the overall performance of the series of tests is similar to that of a single test with a tolerance rate $\lambda_c = 6{\cdot}1$ counts/sec. There is thus a negligible risk that seriously contaminated personnel will pass the test after a reasonably small number of attempts. This is further demonstrated by the case $\lambda = 8{\cdot}77$, for which the probability of acceptance rises from 0·000 54 for one attempt to 0·005 4 for ten attempts.

(10) OTHER APPLICATIONS

The previous Sections have considered mainly hand monitoring, and in particular monitoring for α-particle activity, where the tolerance counting rate is about 5 counts/sec, but there are many other situations where the principles described can be of equal importance.

Much work in such fields as chemistry, health physics, medicine and the application of radioisotopes involves the preparation and measurement of many weak radioactive samples. The counting of these samples can be a lengthy process, and it is often the case that the weakest samples are of little interest, an accurate measurement only needing to be made on the more active ones. Therefore the rapid rejection of weak samples can save a great deal of counting time. To do this, a difference integrator can be used for preliminary sorting of the samples into 'high' and 'low', only the former receiving a subsequent accurate count. In one instance where all samples are at present counted for eight hours, it is proposed to introduce a preliminary sorting operation which has an average sample time of 10 minutes for weak samples. The distribution of activity between samples is such that only about one sample in 100 is expected to require the more accurate eight-hour measurement, and the average time per sample will thus be reduced from eight hours to about 20 minutes. An alternative arrangement is to use the difference integrator in parallel with the normal counting system to give a rapid indication if the sample activity is low (any 'high' indication being ignored). It could be used in this way in conjunction with an automatic counting system incorporating an automatic sample changer to curtail the counting of weak samples.

Section 14.4 shows that the discriminating power of the parallel-boundary test is independent of the value of the tolerance counting rate chosen. The results already obtained can therefore be applied to different situations by choosing the appropriate value of λ_c and adjusting the time scale of the a.s.t. curves in inverse proportion.

The parallel-boundary method does not only provide a means of saving time without losing accuracy: it can also provide increased accuracy without an increase of average sampling time. Use can be made of Fig. 15 to demonstrate this point. The test shown has a discriminating power five times as great as the test of Fig. 9. For the former test λ_c is 25 counts/sec, but for comparison with the latter test, λ_c can be reduced to 5 counts/sec by dividing the counting rates by five and multiplying the time scale by the same factor. The method of Section 14.2 can then be used to calculate the average sampling time for a rectangular distribution of activities stretching from zero to 10 counts/sec. This gives 26·4 sec per test, or if 90% of the samples are uncontaminated, the time is reduced to 8·04 sec. In the corresponding rectangular-boundary test both the count and time limits must be increased 25-fold over the values of Fig. 9 in order to obtain the required fivefold increase of discriminating power. Thus, with the rectangular distribution of samples considered above, the average sampling time is 148 sec, or if 90% of the samples are inactive, the time is increased to 172 sec. It can be con-

cluded that the greater the discriminating power required, the greater is the advantage of using the parallel-boundary test.

The monitoring of surfaces such as floors, benches or clothing is normally carried out by an operator moving a detector over the surface while listening for the audible clicks associated with the detection of radioactive disintegrations. This process is highly subjective, for its sensitivity depends on the speed at which the operator moves the detector and upon his ability to distinguish between random and systematic variations in the pulse rate. Here again the difference integrator could be used to distinguish between significant and insignificant activity. One method of use would be to bring the radiation detector up to the surface to be monitored, keep it stationary during the sampling period, and when a 'low' indication is obtained, move the detector by an amount equal to its width and start a new sampling period. For the same sensitivity this method would be as fast as that now used and would eliminate subjective uncertainties.

This use raises the question whether the sequential sampling principle could be applied to the continuous monitoring of radioactivity so as to give a rapid indication of a change in pulse rate. There are indications that this could be done, but further study is required.

(11) CONCLUSIONS

An efficient qualitative test for determining whether the mean occurrence rate of a random process is above or below a given tolerance level demands the minimum sample size consistent with an acceptable probability of error. Sequential test procedures minimize the sample size by observing the random process until some terminating condition is satisfied. This terminating condition may be a simple function of the cumulative count of events and the duration of the test, chosen so that it defines the probability of error reasonably well and leads to a convenient realization of the test procedure.

The performance of two sequential test procedures has been investigated, with particular reference to the conditions encountered in instruments used for monitoring radioactive contamination on the hands. One procedure is the parallel-boundary test due to Wald, which tests the difference between the number of events actually observed and the number expected at the tolerance rate and terminates the test when this difference exceeds predetermined limits. The other procedure is the rectangular-boundary test, which terminates when either the count of observed events or the duration of the test exceeds predetermined limits. Table 1 summarizes the main numerical results which have been obtained.

It will be seen that the rectangular-boundary test, which has been widely used in hand monitors, is much less efficient than the parallel-boundary test. For a typical distribution of r activity, with a preponderance of uncontaminated personne average duration of the parallel-boundary test (ignoring delay while the radiation detectors and the hands are bro together) is approximately one-fifth of that for the comm used rectangular-boundary test having similar discrimin power.

Although the parallel-boundary procedure greatly reduce average duration of the test, it suffers from the disadva that some samples, with activity near the tolerance level, require an excessively long test before a decision is given. defect can be overcome by imposing a limit to the durati the test. Several alternative methods for applying a time have been examined in sufficient detail to show that they w be satisfactory and relatively simple. Use of the par boundary test appears to introduce little, if any, extra com tion in the electronic circuits; in fact these may in some ins be simplified.

Although the parallel-boundary test has been considered mainly in relation to the advantages of its use for monitoring, its application is to be recommended in almos situation where the activity of many weakly radioactive sa is to be tested.

(12) ACKNOWLEDGMENTS

The authors are grateful to Dr. J. M. Hammersley for tance in the early stages of the work, and in particular for dra their attention to Reference 3, and to Messrs. K. W. Morto J. E. Hailstone for advice on the preparation of the com programme.

Table 1

COMPARISON BETWEEN AVERAGE TIMES TAKEN BY RECTANGULAR-BOUNDARY AND PARALLEL-BOUNDARY TESTS

Conditions of test	Average time for rectangular-boundary test	Average time for equivalent parallel-boundary test
	sec	sec
(*a*) Tolerance rate 5 counts/sec (no background activity). Activity uniformly distributed between zero and 10 counts/sec	5·9	3·35
(*b*) As (*a*), but 90% of samples inactive. Remainder uniformly distributed between zero and 10 counts/sec	6·9	1·41
(*c*) Tolerance rate 5 counts/sec superimposed on 20 counts/sec background. Activity uniformly distributed between background and background plus 10 counts/sec	33·4	16·7
(*d*) As (*c*), but 90% of samples inactive (background only). Remainder uniformly distributed between background and background plus 10 counts/sec	34·8	7·08
(*e*) As (*b*), but discriminating power increased fivefold	172	8·04

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(14) APPENDICES

4.1) Theoretical Derivation of O.C. and A.S.T. Functions for the Parallel-Boundary Test

is required to test the hypothesis that the unknown parar, λ, of a random process with Poisson distribution does exceed a specified value, λ_c. In general it is possible to ose, from purely practical considerations, two values, $< \lambda_c$ and $\lambda_2 > \lambda_c$, such that when $\lambda \leqslant \lambda_1$ the hypothesis uld be accepted with a probability of error not exceeding e small quantity α_1, and when $\lambda \geqslant \lambda_2$ the hypothesis should ejected with a probability of error not exceeding α_2. When s between λ_1 and λ_2 the magnitude of an error is quite small, ne probability of error can be allowed to exceed α_1 and α_2. requirements for the test are thus defined by the four ntities λ_1, α_1; λ_2, α_2.

sequential probability ratio test, of the type developed by d, tests one hypothesis against another. In this case it is ired to test the hypothesis that $\lambda \leqslant \lambda_c$ against the alternative othesis that $\lambda \geqslant \lambda_c$, with tolerated risks of error defined by α_1; λ_2, α_2. It is, however, more convenient to test the othesis that $\lambda = \lambda_1$ against the alternative that $\lambda = \lambda_2$, with s of error again defined by λ_1, α_1; λ_2, α_2. When this latter leads to acceptance of the hypothesis that $\lambda = \lambda_1$ with the rated risk of error, then the risk of error incurred by accept- the hypothesis that $\lambda \leqslant \lambda_c$ is smaller than this, provided λ lies outside the range $\lambda_1 < \lambda < \lambda_2$.

iefer and Wolfowitz[3] consider sequential probability ratio s of this type and give rules of the form:

(*a*) Accept the hypothesis that $\lambda = \lambda_1$ if $n - t\lambda_1/r \leqslant K$.

(*b*) Accept the hypothesis that $\lambda = \lambda_2$ if $n - t\lambda_1/r \geqslant J$.

(*c*) Continue testing if $J > n - t\lambda_1/r > K$

re
$$r = \frac{\log(\lambda_2/\lambda_1)}{(\lambda_2/\lambda_1) - 1}$$
$$K = \frac{\log[\alpha_2/(1-\alpha_1)]}{\log(\lambda_2/\lambda_1)}$$
$$J \leqslant \frac{\log[(1-\alpha_2)/\alpha_1]}{\log(\lambda_2/\lambda_1)} \leqslant J + 1$$

The slope of the boundary lines is thus λ_1/r.

n Section 5 we could, for example, have stated the requirets for the parallel-boundary tests by taking two points on o.c. of the rectangular-boundary test, corresponding to 1% bability of error. Thus $\lambda_1 = 3{\cdot}3$, $\alpha_1 = 0{\cdot}01$, $\lambda_2 = 7{\cdot}1$, $= 0{\cdot}01$, and from the relationships given above we obtain $- 6$, $J \leqslant 6 \leqslant J + 1$, $\lambda_1/r = 4{\cdot}96$. The graphical solution $- K = 6$, slope $= 5$, is thus confirmed.

lternatively, it may be required to tabulate the o.c. and a.s.t. tions for a test with specified values of J and K and for ch the slope of the boundaries is to be λ_c. The o.c. function, , is the probability of accepting the hypothesis that $\lambda \geqslant \lambda_c$.

Reference 3 gives Tables of a function $P(q, J, K)$, where

$$P(q, J, K) = 1 - L(\lambda)$$

and q satisfies the condition $\lambda = \dfrac{\lambda_1 \log q}{r(q-1)}$. In the special case where the slope of the boundaries is λ_c,

$$q = (\lambda_c \log q)/(q - 1)$$

For any required ratio of the mean occurrence rate to the tolerance rate, λ/λ_c, the Tables of P can be entered at the appropriate values of q, J and K to obtain the value of the o.c. function, $L(\lambda)$, and also the average sample number, $h(\lambda)$. The average sample time, $A(\lambda)$, is given by $A(\lambda) = h(\lambda)/\lambda$.

The functions P and h are tabulated in Reference 3 for selected values of J and K in the range 2–10 inclusive. For larger values of J and K the following inequalities, derived from Reference 3 will usually give adequate approximations for $L(\lambda, J, K)$ and $A(\lambda, J, K)$ when the slope of the boundaries is λ_c.

When $\lambda \neq \lambda_c$,

$$\frac{1 - q^K}{q^J - q^K} \leqslant L(\lambda) \leqslant \frac{1 - q^K}{q^{J+1} - q^K} \quad . \quad . \quad . \quad . \quad (1)$$

where $\dfrac{\log q}{q - 1} = \dfrac{\lambda}{\lambda_c}$.

When $\lambda = \lambda_c$,

$$\frac{-K}{J - K} \leqslant L(\lambda_c) \leqslant \frac{-K}{J + 1 - K} \quad . \quad . \quad . \quad . \quad (2)$$

When $\lambda < \lambda_c$,

$$\frac{K + (J - K)L(\lambda)}{\lambda - \lambda_c} \geqslant A(\lambda) \geqslant \frac{K + (J + 1 - K)L(\lambda)}{\lambda - \lambda_c} \quad . \quad (3)$$

When $\lambda > \lambda_c$,

$$\frac{K + (J - K)L(\lambda)}{\lambda - \lambda_c} \leqslant A(\lambda) \leqslant \frac{K + (J + 1 - K)L(\lambda)}{\lambda - \lambda_c} \quad . \quad (4)$$

When $\lambda = \lambda_c$,

$$\frac{K^2 + (J^2 - K^2)L(\lambda_c)}{\lambda_c} \leqslant A(\lambda) \leqslant \frac{K^2 + [(J+1)^2 - K^2]L(\lambda_c)}{\lambda_c}$$
$$. \quad . \quad . \quad . \quad (5)$$

As q is a function of the relative occurrence rate, λ/λ_c, it follows that the o.c. function, $1 - P(q, J, K)$, and the average sample number, $h(q, J, K)$, depend on the relative rate and not on the absolute occurrence rate.

The inequalities given above are not very easily handled and further approximation is desirable. When λ is sufficiently large or small for $L(\lambda)$ to be regarded as equal to 1 or 0, inequality (3) reduces to

$$A(\lambda) = K/(\lambda - \lambda_c)$$

and inequality (4) reduces to

$$\frac{J}{\lambda - \lambda_c} \leqslant A(\lambda) \leqslant \frac{J + 1}{\lambda - \lambda_c}$$

For the cases considered in the paper the computed results show that when $L(\lambda)$ tends to unity, $A(\lambda)$ is very nearly equal to $J/(\lambda - \lambda_c)$. When $\lambda = \lambda_c$, then $L(\lambda)$ is approximately $\frac{1}{2}$, and if, in addition, it is assumed that $J = -K$, then inequality (5) reduces to

$$\frac{J^2}{\lambda_c} \leqslant A(\lambda) \leqslant \frac{J^2 + J + \frac{1}{2}}{\lambda_c}$$

In the cases computed in the paper, the a.s.t. at tolerance counting rate tends towards the lower of these two limits.

(14.2) Determination of Average Sampling Time of a Parallel-Boundary System counting a Rectangular Distribution of Activities

Section 14.1 shows how the average sampling time for a given value of radioactivity can be ascertained. To obtain the average sampling time for a rectangular distribution of activities it is necessary to calculate the area under the a.s.t. curve between specified values of the true counting rate. It is difficult to obtain an exact analytical expression for this curve, so the approximations given in Section 14.1 will be used as follows:

When

$$\lambda < \left(\lambda_c + \frac{\lambda_c}{K}\right), \qquad A(\lambda) = -K/(\lambda_c - \lambda)$$

When

$$\left(\lambda_c + \frac{\lambda_c}{K}\right) < \lambda < \left(\lambda_c + \frac{\lambda_c}{J}\right) \text{ and } J = -K, \qquad A(\lambda) = J^2/\lambda_c$$

When

$$\lambda > \left(\lambda_c + \frac{\lambda_c}{J}\right), \qquad A(\lambda) = J/(\lambda - \lambda_c)$$

The broken curve in Fig. 8 shows how the a.s.t. curve obtained from these approximations compares with an a.s.t. curve obtained as described in the paper. It will be seen that the areas under the two curves differ only slightly and that, if anything, the approximation tends to overestimate the area. Using this approximation, the area under the curve between two counting rates, λ_a and λ_b, when $J = -K$, is given by

$$\int_{\lambda_a}^{\lambda_c+\lambda_c/K} -\frac{K}{\lambda_c - \lambda} d\lambda + \frac{J^2}{\lambda_c}\left(\frac{\lambda_c}{J} - \frac{\lambda_c}{K}\right) + \int_{\lambda_c+\lambda_c/J}^{\lambda_b} \frac{J}{\lambda - \lambda_c} d\lambda$$

This is equal to

$$J\left(2 + \log_\varepsilon J\frac{\lambda_c - \lambda_a}{\lambda_c} + \log_\varepsilon J\frac{\lambda_b - \lambda_c}{\lambda_c}\right)$$

To obtain the average sampling time, this expression is divided by the length, $\lambda_b - \lambda_a$, of the rectangular distribution, so the average sampling time for this distribution is

$$\frac{J}{\lambda_b - \lambda_a}\left(2 + \log_\varepsilon J\frac{\lambda_c - \lambda_a}{\lambda_c} + \log_\varepsilon J\frac{\lambda_b - \lambda_c}{\lambda_c}\right)$$

When, as in the case considered in the paper, the rectangular distribution extends from $\lambda = 0$ to $\lambda = 2\lambda_c$, then $\lambda_a = 0$ and $\lambda_b = 2\lambda_c$. The above expression then reduces to

$$J(1 + \log_\varepsilon J)/\lambda_c$$

(14.3) Determination of the Average Sampling Time o Rectangular-Boundary System counting a Rectang Distribution of Activities

As in Section 14.2, it is convenient to derive a simple appr mate expression for the a.s.t. function. Counting rates nea are counted with a standard deviation of approximately $\lambda/$ so that, when $\lambda < \lambda_c - \lambda_c/\sqrt{n_b}$, we can suppose that r samples are time-limited and the a.s.t. is t_b. Likewise, w $\lambda > \lambda_c + \lambda_c/\sqrt{n_b}$ most samples are count-limited and the a is n_b/λ.

When $\lambda_c - \lambda_c/\sqrt{n} < \lambda < \lambda_c + \lambda_c/\sqrt{n_b}$, some samples time-limited and some are count-limited. The broken cur Fig. 5 shows that in this region a good approximation is achi by linear interpolation. Thus, the area under the a.s.t. c between $\lambda = \lambda_a$ and $\lambda = \lambda_b$ is

$$t_b\left(\lambda_c - \frac{\lambda_c}{\sqrt{n_b}} - \lambda_a\right) + \left(t_b + \frac{1}{2}\frac{n_b}{\lambda_c + \lambda_c/\sqrt{n_b}}\right)2\frac{\lambda_c}{\sqrt{n_b}} + \int_{\lambda_c+\lambda_c}^{\lambda_b} \frac{n_b}{\lambda}$$

By definition, $t_b = n_b/\lambda_c$, so the expression reduces to

$$n_b\left[1 - \frac{\lambda_a}{\lambda_c} + \frac{1}{1 + \sqrt{n_b}} + \log_\varepsilon\left(\frac{\lambda_b}{\lambda_c + \lambda_c/\sqrt{n_b}}\right)\right]$$

The average sample time between λ_1 and λ_2 is therefore

$$\frac{n_b}{\lambda_b - \lambda_a}\left[1 - \frac{\lambda_a}{\lambda_c} + \frac{1}{1 + \sqrt{n_b}} + \log_\varepsilon\left(\frac{\lambda_b}{\lambda_c + \lambda_c/\sqrt{n_b}}\right)\right]$$

The rectangular distribution discussed in the paper extends f $\lambda = 0$ to $\lambda = 2\lambda_c$, so that $\lambda_a = 0$ and $\lambda_b = 2\lambda_c$ and the expres reduces to

$$\frac{n_b}{2\lambda_c}\left[1 + \frac{1}{1 + \sqrt{n_b}} + \log_\varepsilon\left(\frac{2}{1 + 1/\sqrt{n_b}}\right)\right]$$

(14.4) Discriminating Power of Parallel-Boundary Test

For given values of α_1 and α_2 (see Section 14.1), the p of the test to discriminate between $\lambda > \lambda_c$ and $\lambda < \lambda_c$ ma expressed as $\lambda_c/(\lambda_2 - \lambda_1)$.

Taking λ_c as the slope of the boundary lines (see Section 1

$$\lambda_c = \frac{\lambda_1}{r} = \frac{\lambda_2 - \lambda_1}{\log(\lambda_2/\lambda_1)}$$

Thus, for probabilities of error α_1 and α_2, the discrimin power D is $1/\log(\lambda_2/\lambda_1)$. If $J = -K$ and $\alpha_1 = \alpha_2 = \alpha$, $J = -K = D\log[(1-\alpha)/\alpha]$ and therefore $D = J/\log[(1-\alpha$

If $\alpha = 0{\cdot}27$, $D = J$ or $-K$.

If $\alpha = 0{\cdot}01$, $D = 0{\cdot}218(J$ or $-K)$.

DISCUSSION BEFORE THE MEASUREMENT AND CONTROL SECTION, 25TH OCTOBER, 196

Mr. B. D. Corbett: The authors have enunciated an elegant principle which should obviously be applied in all future instruments designed for checking radioactive contamination of the hands. The need to save time in such tests is particularly acute at nuclear power stations where a large number of men, all anxious to get home, come off shift at the same time.

The authors have analysed the function of an ideal difference integrator, which may be considered as a pulse ratemeter having an infinitely long integrating time-constant. I have given some consideration to the application of their principle in a circuit having a finite integrating time-constant, i.e. the capacitor has a leak resistance connected across it. This was prompted wish to use simple transistorized circuits. Using the aut symbols, we have, for the integrator or countmeter circuit,

$$\text{Coefficient of variation} = 1/\sqrt{(\lambda t)} \text{ at time } t.$$

For a ratemeter circuit with integrating time-constant, τ,

$$\text{Coefficient of variation} = \frac{1}{\sqrt{(2\lambda\tau \tanh t/2\tau)}}$$

This approaches the limiting value $1/\sqrt{(2\lambda\tau)}$ when $t \gg \tau$.

Hence it may be shown that the loss of accuracy of the

as compared with the countmeter circuit is small pro- that the time required to reach a trip point is equal to or han the integrating time-constant. Table A gives the

Table A

t/τ	Increase in assessment time
	%
0·1	0·08
0·5	2·0
1·0	7·6
1·5	15·3

ve increase of time required by the ratemeter for a given tical accuracy. The authors' statistics may therefore be ed in the design for a ratemeter circuit using the parallel- dary sequential testing principle.

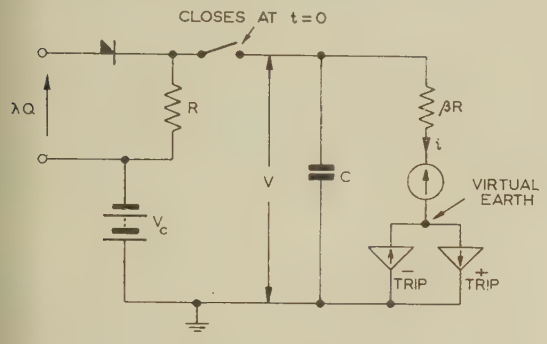


Fig. A.—Biased ratemeter circuit.

Q = Charge per pulse.
V = Tank voltage.
V_c = Reference bias voltage.
i = Metering current.
R, βR = Leak resistors.
Time-constant is $\tau = \frac{\beta}{1+\beta} RC$

nsider the biased ratemeter circuit shown in Fig. A. At brium,

$$\lambda Q = \frac{V}{\beta R} + \frac{V + V_c}{R}$$

ence

$$i = \frac{\lambda Q - V_c/R}{1 + \beta}$$

neral, one or other of the trip circuits will operate before brium is reached and the following equation applies:

$$i = \frac{\lambda_c Q}{1 + \beta}(m - 1)(1 - \varepsilon^{-t/\tau})$$

m is the number of maximum permissible levels and s λ/λ_c. Background can be subtracted by control of V_c subtracting pulses from the common tank through another (Fig. B). The type of ratemeter envisaged in these circuits

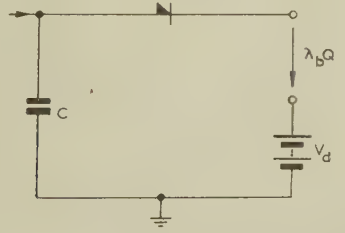


Fig. B.—Background subtraction by pulses.

$$V_d \geqslant \frac{\beta}{1+\beta} V_c$$

is one using inductive storage of energy, enabling voltages of the order of 100 volts or more to be used on the ratemeter integrating capacitor.

The authors' principle suggests the need for double-input reversible scalers to enable the difference-integration process to be carried out digitally. Is this being considered? Further, is an indication of the expected value of the detected activity desirable in some cases? This could be provided by a suitably calibrated clock to time the process in conjunction with the high and low indicators.

Dr. Denis Taylor: For some years I have been conscious of the need to economize in counting time and in a book* published in 1958, I mentioned the use of a preliminary rough assay, and was interested in methods of the type described in the paper.

The application to hand and to filter-paper monitoring seemed quite definite. I wonder whether we could not apply the principle to problems connected with burst-fuel-element detection systems. One problem has been to devise a satisfactory system with which to operate when a reactor is shut down. One solution involves monitoring α-activity from the oxide type of fuel. The activities may be very low, and so the special conditions considered by the authors apply here. However, although there are a large number of fuel-element channels to monitor in a power reactor, one does not necessarily need to complete the cycle in an extremely short time if it is shut down. If a cheaper and simpler apparatus involving a longer cycle time can be used, it is probably still acceptable.

Another important field is body monitoring. One might be faced with a situation where a relatively large number of people had been involved in an accident following the escape of radioactive material. The initial problem would be to sort out those who had been contaminated by inhaling or absorbing radioactivity. The matter has to be dealt with very quickly, and this might be another case where the authors' method would be of value.

Could the technique have a wider application beyond the field of health physics? Many of us are interested in process and plant control and perhaps the parallel-boundary technique could be applied there. One is concerned with keeping the activity between two set limits rather than deciding whether it is above or below a predetermined level. There are a number of processes where the instrumentation is used to indicate whether the control system is working properly and the activity is between preset limits. One can think of having something of this kind over and above the control system as a quick-acting mechanism to give warning if the activity is outside the particular limits necessary for proper control of the process, and it does not seem to represent such a tremendous step from what we have heard to-night. Have the authors considered the problem and might their techniques be applicable to it?

Mr. R. D. Trotter: The authors have transmuted the statement that '35 counts or more in seven seconds is about tolerance level' into the following:

> (*a*) If the test has continued for a predetermined time, t_b, reject the hypothesis that the observed occurence rate is greater than the tolerance rate and terminate the test.
> (*b*) If the cumulative total of pulses has reached a predetermined value, n_b, accept the hypothesis and terminate the test.

That sort of thing frightens people away from statistics, even though a more scientific application of logic is no doubt very good for everyone concerned.

Fig. 9 is the key diagram, especially curves (*b*) and (*d*), which show the substantial saving in time for a large number of tests which can be made by using the parallel-boundary method when

* Taylor, D.: 'Radiation Hazards and Protection' (Newnes, 1958), p. 142.

the signal level is well away from a critical value. The curves drawn are for operating characteristics which are similar for the two techniques, because health physicists cannot make up their minds on the real criteria for decision. It seeems to me that the accuracy of the pulse-height definition is an equal hazard to either method, but in the case of the parallel-boundary technique the stability of the current leakage of the difference integrator must be beyond reproach and not introduce a further source of error.

I also am interested in the possibility of the technique beng applied to burst fuel-element detection systems, but I have more in mind burst-slug detection when the reactor is operating at power. Here the problem is to sample many channels seq ally, inspecting each in turn. Because it is necessary to go the whole reactor in a reasonable time, with a reasonable a of equipment, the time one that can spend on each indi sample is limited. Can parallel-boundary techniques in the operating characteristic that is obtained for a sam shorter time than one would really like for a statistical The problem is to detect when the signal is above the tol level, with a sample that has large random fluctuations. I any possibility that these techniques can also be used in a way—to provide us with the rate of change of activity de between reading or sampling times of very short duration

THE AUTHORS' REPLY TO THE ABOVE DISCUSSION

Messrs E. H. Cooke-Yarborough and **R. C. M. Barnes** (*in reply*): Mr. Corbett's calculation of the behaviour of a difference integrator circuit with a finite time-constant is a useful extension to our paper. We investigated this case to the extent of showing that the average sample time was considerably lengthened at counting rates close to λ_c. We went no farther, because we found that a transistor circuit giving an adequately long time-constant could be devised quite simply. Fig. C shows the

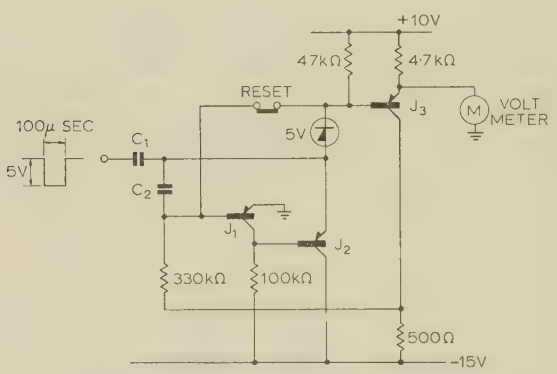


Fig. C.—Circuit for a difference integrator.

C_1 2μF C_2 10μF J_1, J_3 Type OC201 J_2 Type XA102

circuit used in the equipment demonstrated. Transistors J_1 and J_2 form, with C_2, a Miller integrator. They also function as the two diodes of a diode pump circuit. The negative-going input pulse cuts off J_2 emitter and turns J_1 on, and C_1 and C_2 then discharge in series through the base of J_1. This changes the voltage across C_2 by 1/6th of the input pulse voltage. At the end of the pulse J_1 cuts off and C_1 recharges through J_2 emitter. When this process is complete the normal Miller feedback is restored. The 5-volt avalanche diode provides the voltage displacement necessary for the meter to read zero at the centre of the range of the integrator. When the reset contact is closed the bases of J_1 and J_3 are at the same potential, and since both transistors are of the same type and J_1 emitter is earthed, the emitter potential of J_3 is very close to earth and zero drifts are largely eliminated.

The Miller integrator has finite loop gain, so the time-constant is also finite, but it can be substantially increased by positive feedback applied from the collector of J_3 to the low-potential end of the difference-integrator leak-resistor. A time-constant long compared with the sampling time can be achieved without critical adjustment.

A difference integrator using a reversible scaler has been described by Moss.* So long as J and K can be kept wit limits of a single Dekatron, the complications of rev carry are avoided and the circuits present no difficulty.

A suitably calibrated clock could certainly allow an ap mate indication of the strength of the activity, but co might arise from the double-valued readings which wo obtained. A better arrangement might be that sugges the penultimate sentence of the second paragraph of Sect of the paper. Here, if a time limit is applied, the scaler g direct measure of the counting rate if this exceeds λ_c, wh lower counting rates the contents of the scaler are a non but non-ambiguous function of the counting rate.

In reply to Dr. Taylor, the principle described shou applicable to burst-fuel-element detection systems in situ where the limitation on detection is due to counting sta and where the mechanism for sampling different chan such that the sampling times can be varied. The system be applicable to continuous monitoring of activities circuits are arranged so that the difference integrator i as soon as an indication is obtained. We have briefly inves continuous monitoring by use of a difference integrator appears that, while large changes of activity give a w sooner with the difference integrator than with a conve counting rate meter and trip circuit, the reverse is true change is small.

While we concede that the two sentences to which Mr. has taken exception might have been less formally word cannot agree that they are equivalent to his statement '35 or more in seven seconds is about tolerance level', wh neither general nor rigorous. They are preceded by two of less-rigorous description of the two procedures. The expressed view that 'anything can be proved with sta usually results from loose descriptions of the procedures in On the other hand, it is difficult to be rigorous without wording which is almost legal in character.

The difference integrator is no more sensitive to leakage than a conventional ratemeter. In either case a percentage change in the effective value of the leak resist produce the same percentage change in the apparent value

The parallel-boundary test cannot, in a given length o obtain information about an individual sample which is s cally more significant than that obtained by any other t it is of no advantage if the time available to examine an ind sample is fixed. If, however, it is only the overall time to e many samples which is limited, and the time for ind samples can be varied, then the advantages of the p boundary test can be exploited. We have not inves techniques for detecting rate of change of activity, but it likely that a similar rapid method exists.

* Moss, G. H.: 'A Medical Application of Difference Integration', Proce the Third International Conference on Medical Electronics, London, 196 published).

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DISCRETE ANALOGUE-COMPUTER COMPENSATION OF SAMPLED-DATA CONTROL SYSTEMS

By T. GLUCHAROFF, M.E.

aper was first received 25th February, and in revised form 23rd July, 1960. It was published in November, 1960, and was read before the MEASUREMENT AND CONTROL SECTION *15th November, 1960.)*

SUMMARY

paper describes an analogue computer which can solve pulse er functions and operate as a discrete controller for compensa- f sampled-data or continuous systems. Operational amplifiers licon-diode switches are combined to perform the basic functions ıpling, holding and time delay. The accuracy obtainable is com- le with that of digital controllers, but the analogue computer e advantages of simplicity of setting-up and low cost. Adjust- of the controller parameters is easy, and the computer can be used ermine the coefficients of the pulse transfer function for optimum tion before a digital computer is employed. A simplified method crete-controller design for saturating sampled-data systems is resented.

LIST OF SYMBOLS

c, d = Constant or gain of a computing amplifier.
n, q = Integer.
p = Laplacian operator.
v = Circuit input voltage.
v' = Amplifier input voltage.
v'' = Amplifier output voltage.
v_i = Diode switch or transistor switch input voltage.
$z = \varepsilon^{pT}$.
δ = Sampling interval.
$c(t)$ = Output time function.
$h(t)$ = Response of the plant to a unit pulse input.
$H(z)$ = Pulse sequence of $h(t)$.
$r(t)$ = Input time function.
$w(t)$ = Response of the plant to a unit step input.
$W(z)$ = Pulse sequence of $w(t)$.
C_f = Feedback computing capacitance of an operational amplifier.
K = Amplifier gain.
R_b = Base resistance in a transistor switch.
R_d = Incremental resistance of a silicon diode.
R_h = Feedback computing resistance of an operational amplifier.
R_i = Input computing resistance of an operational amplifier.
R_p = Switching-voltage resistance of a diode switch.
T = Sampling period.
$G(p)$ = Transfer function of the plant or that of a system.
$C(z)$ = Pulse sequence of the output of a system.
$D(z)$ = Pulse transfer function of the discrete controller.
$E(z)$ = Pulse sequence of the error function.
$G(z)$ = Pulse transfer function of a system.
$F(z)$ = Pulse sequence of the forcing function.
$R(z)$ = Pulse sequence of the input to a system.

Glucharoff is in the School of Electrical Engineering, University of New South Australia. At the meeting the paper was read on his behalf by Dr. J. H. t.

(1) INTRODUCTION

(1.1) Sampled-Data Systems

During the last decade sampled-data systems have become of considerable interest and importance to engineers and scientists. In such systems signals appear at one or more points of the system, not as continuous but as discrete functions of the time variable t. These signals are usually in the form of pulses or numbers applied at equally spaced intervals of time, and no information is received between two consecutive signals. For example, if a digital computer is used, the data input and output are applied intermittently, and its operation is essentially that of a sampled-data system. The rapid advancement in the field of digital computers, and their increased reliability from the use of solid-state components, have made the application of such computers in large-scale systems practical. There are numerous other examples of sampled signals in control systems; the sampled information may be inherent to the type of system, or it is purposely introduced where the use of sampled-data techniques results in a control system superior to an alternative unsampled system.

The behaviour of sampled-data systems is most easily studied by means of the z transform,[1,3,4] an operational method similar to the Laplace transform for continuous systems. The z transform can be considered as a modification of the Laplace transform or approached directly as the operational calculus of number sequences. In the following, a simplified approach has been adopted to illustrate some of the properties of the z transform without mathematical rigour.

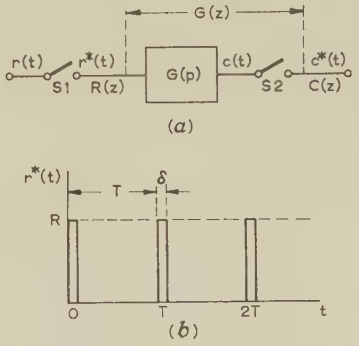


Fig. 1.—Sampled-data system.
(*a*) Linear system between two samplers.
(*b*) Sampler output for a step input.

Consider the linear system $G(p)$ in Fig. 1(*a*), with a sampler at both its input and output. The samplers are represented by periodic switches, which are closed during the short sampling intervals δ and open for the rest of the periods T, so that, for a step input $r(t) = R$ the output of the sampler S1 is as shown in Fig. 1(*b*). If the sampling interval δ is sufficiently small the

pulses can be considered to be impulses of area R, and the Laplace transform of the sampled input $r^*(t)$ becomes

$$R^*(p) = R(1 + \varepsilon^{-pT} + \varepsilon^{-2pT} + \varepsilon^{-3pT} + \ldots)$$

If a change of variable is made such that $z = \varepsilon^{pT}$ the transform of the sampled sequence can be expressed in terms of the new variable z as

$$R(z) = R(1 + z^{-1} + z^{-2} + z^{-3} + \ldots) = \sum_{k=0}^{\infty} Rz^{-k}$$

This infinite summation is called the pulse sequence of the input signal, and $z^{-1} = \varepsilon^{-pT}$ represents a time delay of one sampling period T. When the input signal $r(t)$ is some other function of time, the output sequence of the sampler is simply given by

$$R(z) = \sum_{k=0}^{\infty} r(kT)z^{-k}$$

where $r(kT)$ is the value of $r(t)$ at $t = kT$.

If the Laplace-transform $R(p)$ of the input time function $r(t)$ is a ratio of finite polynomials in p, the pulse sequence of the sampled function can be expressed in closed form as the ratio of finite polynomials in z, and is called the z transform or the pulse transform of the time function. The z transform can be calculated by well-known mathematical methods and is extensively tabulated.[1,4] For the above simple example, however, when $r(t)$ is a step input, the z transform can easily be determined directly from the pulse sequence as

$$R(z) = R(1 + z^{-1} + z^{-2} + z^{-3} + \ldots) = \frac{R}{1 - z^{-1}}$$

Conversely, when the z transform is available the pulse sequence can always be obtained by long division; for example, dividing $(1 - z^{-1})$ into R yields the above pulse sequence for $R(z)$.

Consider now the output of the sampler S2 of the system in Fig. 1(*a*). The z transform $C(z)$ of this output differs from the z transform $R(z)$ of the output of the sampler S1, and the manner in which the pulse sequence has been modified by $G(p)$ is given by

$$G(z) = \frac{C(z)}{R(z)}$$

This is called the pulse transfer function or the z transfer function of $G(p)$, and can be calculated in the same manner as the z transform of a time function given by its Laplace transform $R(p)$. For a system whose Laplace transform $G(p)$ has a finite number of poles, the z transform is a ratio of finite polynomials in z, and gives the pulse transfer function as

$$G(z) = \frac{a_0 + a_1z^{-1} + a_2z^{-2} + \ldots + a_nz^{-n}}{1 + b_1z^{-1} + b_2z^{-2} + \ldots + a_mz^{-m}}$$

$$= a_0\frac{(1 - c_1z^{-1})(1 - c_2z^{-1}) \ldots (1 - c_nz^{-1})}{(1 - d_1z^{-1})(1 - d_2z^{-1}) \ldots (1 - d_mz^{-1})}$$

where the coefficients a_0, a_1, a_2 . . . and b_1, b_2 . . . are real, and c_1, c_2 . . . and d_1, d_2 . . . can be real or complex and are the zeros and the poles, respectively, of $G(z)$ in the z-plane. To determine whether a system with a pulse transfer function $G(z)$ is stable or unstable, the locations of its poles in the z-plane are examined in a similar manner as the stability of a continuous system would be determined. When a system transfer function $G(p)$ has poles in the left half of the p-plane only, the system is stable [Fig. 2(*a*)], and the limiting condition is $p = j\omega$, i.e. the imaginary axis. To determine the stability of a system given by a pulse transfer function, the same limiting condition applies

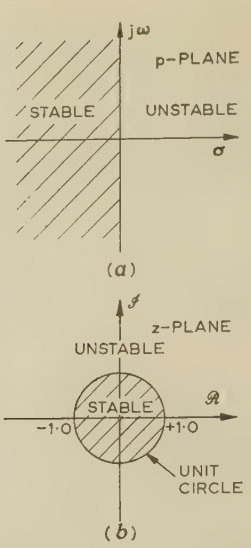


Fig. 2.—Regions of stable and unstable poles.
(*a*) Continuous systems.
(*b*) Sampled-data systems.

in the p-plane, but in the z-plane this condition becomes $z =$ i.e. the unit circle. Therefore, pulse transfer functions z poles inside the unit circle only represent stable sys [Fig. 2(*b*)], and pulse transfer functions with z poles outsi unit circle represent unstable systems.

It can be shown that z functions may be manipulated braically in the same manner as p functions, and that principle of superposition applies, i.e. sampled-data system linear. Hence, the z transform technique can be used to an not only open-loop but also closed-loop sampled-data syst and the stability condition indicated above is directly appli to the overall pulse transfer function of a feedback system. noted, however, that the z transform considers all signals a sampling instants only, and that a modified z transform or technique must be used to determine the behaviour of the si between sampling instants.

Sampling can be considered to be a modulation process, usually, before a sampled signal can be utilized in a u manner, a restoration of the continuous function is requ This can be achieved by the use of a low-pass filter, but a b method is to employ a holding circuit or data-reconstru device. A holding circuit holds its output at the value o

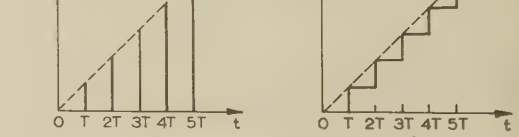


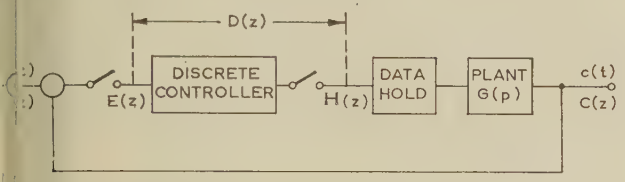
Fig. 3.—Data reconstruction by a holding circuit.
(*a*) Input to the holding circuit.
(*b*) Output of the holding circuit.

previous sample, so that the sampled ramp input in Fig. converted into the staircase function shown in Fig. 3(*b*), is a much better approximation to the continuous functio is noted that the output function is equal to the input fun at sampling instants, and therefore the z transform of the o of a holding circuit is the same as that of its input. S

ng circuit would normally be employed in a sampled-data ack control system, to reconstruct the continuous function the sampled signal, before it was applied to the plant to be olled.

(1.2) Discrete Controllers in Feedback Systems

mpled-data feedback control systems are usually error-led, and a discrete controller $D(z)$ can be used to com-ate the plant to be controlled $G(p)$, as shown in Fig. 4.



ig. 4.—Error-sampled feedback control system with discrete controller.

discrete controller is described by a pulse transfer function, purpose of its introduction being to force satisfactory feed-control. The input $r(t)$ and the output $c(t)$ of the system continuous functions of time, and $R(z)$ and $C(z)$ are the nsforms of these functions. The desired pulse transfer tion $D(z)$ is of the form

$$D(z) = \frac{a_0 + a_1 z^{-1} + a_2 z^{-2} + \ldots a_n z^{-n}}{1 + b_1 z^{-1} + b_2 z^{-2} + \ldots a_m z^{-m}}$$

it is physically realizable only if the denominator contains erm in z^0. This is required because the expansion of $D(z)$ a power series must not contain terms to positive powers It is noted that the numerator of $D(z)$ can contain z^{-1} to power and that often terms like a_0 or a_1 are zero.

iscrete controllers can be divided into several broad groups:

1) *Pure digital computers and computers with digital storage analogue arithmetic operations.*—Such controllers require al-to-analogue and analogue-to-digital convertors, digital ge on drums, tapes or cores and the associated control nits. The accuracy of such controllers is limited to that of convertors and seldom exceeds 0·1%. Special-purpose puters may be of moderate cost and time sharing is possible, hat a number of feedback systems can be compensated with same computer. In large-scale systems the computer may, he same time, be used to perform other functions such as logging, which is widely used at present. In general, the of a digital computer to compensate a single system would uneconomical, and also the adjustment of the controller meters for optimum operation may be very difficult unless plant transfer function is known with high accuracy.

) *Pulsed filters.*[5]—There is no difficulty in implementing controllers, but they are not as general as digital computers. parameters to be adjusted are not directly related to the ficients of the pulse transfer function, and also a discrete ntity is converted into a continuous quantity in order to rate on it, and again it is converted into a discrete quantity he output. It is obvious that this is not a direct method of lementing a pulse transfer function.

) *Pulsed analogue computers.*—In such computers all quan-s appear only in discrete form, and storage and arithmetic rations are performed in an analogue manner. The pro-mming is similar to that of a digital computer, except that various coefficients are set by means of potentiometers or stors.

nalogue computers have the advantages of simplicity in orming the required arithmetic operations, easy program-ming and low cost. The basic principles of operation of pulsed analogue computers are described in the following Section, together with programming methods for such computers.

(2) PULSED ANALOGUE COMPUTERS

Consider a physically realizable pulse transfer function $D(z)$, split into partial fractions

$$D(z) = \frac{a_0 + a_1 z^{-1} + a_2 z^{-2} + \ldots}{1 + b_1 z^{-1} + b_2 z^{-2} + \ldots} = \frac{c_1}{1 - d_1 z^{-1}} + \frac{c_2}{1 - d_2 z^{-1}} + \ldots + \frac{c_r}{1 - d_r z^{-1}} + \ldots$$

$$= D_1(z) + D_2(z) + \ldots + D_r(z) + \ldots$$

and assume that all coefficients $c_1, c_2 \ldots$ and $d_1, d_2 \ldots$ are real. It is seen that if a pulse transfer function of the form $D_r(z) = c_r/(1 - d_r z^{-1})$ can be realized in an analogue manner, the total transfer function $D(z)$ is obtainable by simple addition.

To realize $D_r(z)$, one writes

$$D_r(z) = \frac{C_r(z)}{R(z)} = \frac{c_r}{1 - d_r z^{-1}}$$

$$C_r(z) - C_r(z) d_r z^{-1} = c_r R(z)$$

$$C_r(z) = c_r R(z) + d_r C_r(z) z^{-1}$$

and the corresponding analogue representation is shown in Fig. 5.

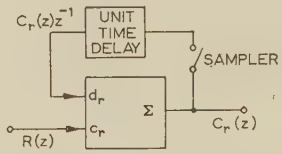


Fig. 5.—Analogue realization of the pulse transfer function

$$D_r(z) = \frac{C_r(z)}{R(z)} = \frac{c_r}{1 - d_r z^{-1}}$$

It is evident that one unit time delay, i.e. a time delay of one sampling period, one addition and a sampler are required to realize any one of the simple transfer functions $D_r(z)$; an additional inversion may be required to obtain the polarity associated with the algebraic sign of d_r. The input and the output are considered at the sampling instants only, and it is noted that the sampler in the feedback path ensures that signals circulating in the feedback loop are discrete functions of the time variable t. This sampler can be located in the feedback path or in the forward path of the feedback loop; in the electronic analogue computer described in this paper a sampler is incorporated in each unit time delay, and therefore the simple transfer function $D_r(z)$ can be realized with one such unit time delay and summing operational amplifiers. Normally, all samplers are omitted from the diagram of a computer arrangement, and also the sampler which follows the arrangement of the total transfer function $D(z)$ is not shown.

The above method of setting up the computer is known as the parallel-programming or partial-fraction method, and uses the minimum number of unit time delays. The required minimum number of unit time delays is equal to the number of poles of the pulse transfer function. It may be noted that, in addition to the real poles considered above, a pulse transfer function may also have poles at the origin and complex poles. Poles at the origin are always present when the order of the nominator

polynomial in z^{-1} exceeds the order of the denominator polynomial in z^{-1}. A pole at the origin represents a time delay of one sampling period and is simply realized by means of a unit time delay. Complex poles always occur in conjugate pairs and may be combined into a single transfer function with real coefficients; such a pair of poles is usually realized by the more general method of programming described below.

A more general method of programming which permits the setting up of the computer parameters directly in terms of the coefficients of a pulse transfer function, and also uses the minimum number of unit time delays, has been developed by Barker.[1] Consider now the pulse transfer function $D(z)$ divided into two factors

$$D(z) = \frac{a_0 + a_1 z^{-1} + a_2 z^{-2} + \ldots}{1 + b_1 z^{-1} + b_2 z^{-2} + \ldots}$$

$$= \left(\frac{1}{1 + b_1 z^{-1} + b_2 z^{-2} + \ldots}\right)$$

$$(a_0 + a_1 z^{-1} + a_2 z^{-2} + \ldots) = D_1(z) D_2(z)$$

The block diagram for $D_1(z)$ is shown in Fig. 6, where each unit time delay is represented by a square box and its gain by a number or symbol inside the box.

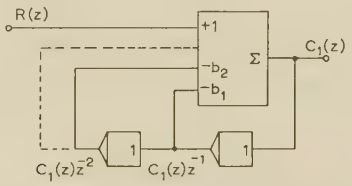


Fig. 6.—Block diagram of the pulse transfer function

$$D_1(z) = \frac{C_1(z)}{R(z)}$$

From Fig. 6 we have

$$C_1(z) = R(z) - C_1(z)(b_1 z^{-1} + b_2 z^{-2} + \ldots)$$

or

$$\frac{C_1(z)}{R(z)} = \frac{1}{1 + b_1 z^{-1} + b_2 z^{-2} + \ldots} = D_1(z)$$

The total output $C(z)$ is obtained by operating on $C_1(z)$ by $D_2(z)$, giving

$$C(z) = C_1(z) D_2(z) = C_1(z)(a_0 + a_1 z^{-1} + a_2 z^{-2} + \ldots)$$

but, since $C_1(z)$, $C_1(z)z^{-1}$, $C_1(z)z^{-2}$, . . . are already available, only one additional summation is required, as shown in Fig. 7.

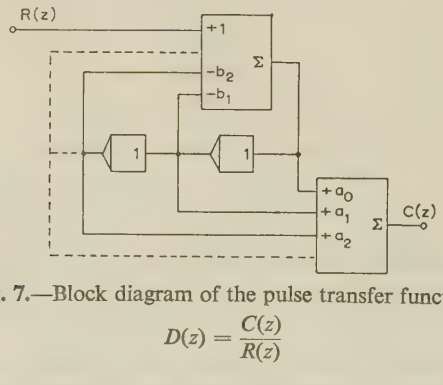


Fig. 7.—Block diagram of the pulse transfer function

$$D(z) = \frac{C(z)}{R(z)}$$

It is noted, however, that another inverting amplifier may be required to obtain the correct polarity according to the algebraic sign of the coefficients.

The necessary time delay in the above computer arrange which is equivalent to storage in digital computers, has bee main obstacle to the development of discrete analogue puters. Electric and acoustic delay lines are suitable onl time delays shorter than 1 millisec, but control systems require sampling periods of many seconds. Magnetic tap drums have been used to obtain time delays, but this me has not been fully investigated and the necessary mecha drive is not a desirable feature of this type of analogue sto

In the analogue computer described in the paper, time d are obtained by means of two sampler-hold circuits operati equal but time-shifted sampling periods. Only operat amplifiers are employed and time delays ranging from a milliseconds to many seconds can be realized. In the follo Section the basic elements of the computer are describe their performances are assessed.

(3) BASIC COMPUTER ELEMENTS

(3.1) The Sampler

The function of the sampler is to convert continuou into a sequence of pulses, and it can take the form of a sw which is closed during the sampling interval δ and is ope the rest of the sampling period T [Fig. 1(*b*)]. Diode gate be used, but such a sampler must work into a high-impe network with good zero stability and frequently an oper amplifier follows the gate.[6] However, a diode gate opera a switch and an operational amplifier can be combine circuit as shown in Fig. 8, with a number of advantages.

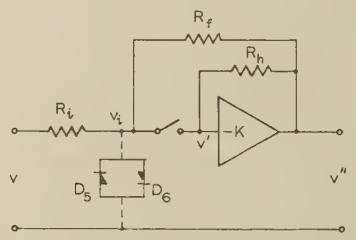


Fig. 8.—Circuit diagram of the sampler.

For an ideal switch and in the presence of an input volt the output voltage v'' is very nearly

$$v'' = -v \frac{R_f R_h / (R_f + R_h)}{R_i}$$

when the switch is closed and $v'' = 0$ when the switch is $v_i = v'$ being very small for a large amplifier gain K.

Consider now the ideal switch replaced by the 4-diode s of Fig. 9(*a*), and its equivalent circuit when the switch is in Fig. 9(*b*). This introduces series and parallel resistan the input of the amplifier, which are due to the increr resistances of the diodes, R_d, and the switching-voltage resist R_p, respectively, and their effects can easily be determi desired.[7] It is sufficient here to state that the effect parallel resistance R_p will be small if the amplifier gain is and that the effect of R_d will be negligible if R_h is chosen large enough. Under these conditions the gain of the sa depends only upon R_i, R_f and R_h, as for the ideal s Another effect of the parallel resistance R_p is to increase th put due to drift in the operational amplifier, and for this r R_p must be kept large; the use of a drift-corrected amplifi effectively eliminate this effect.

At this point it is pertinent to discuss the switching v required to open and close the diode switch. The diode s

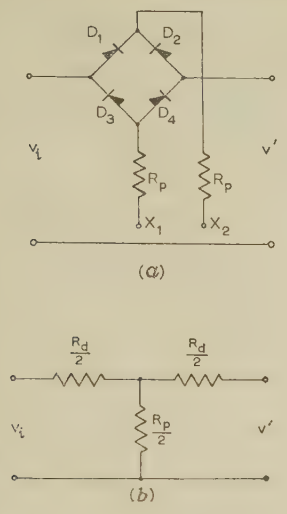


Fig. 9.—Four-diode switch.
(*a*) Circuit diagram.
(*b*) Equivalent circuit applicable only when the switch is closed.

pen whenever the terminals x_1 and x_2 are at zero potential. is the case because a positive voltage v_i due to an input v pass through D_1 but is blocked by D_2. Similarly, a negative age v_i will pass through D_3 but is blocked by D_4. In cases re output due to leakage in the diodes is undesirable, two itional silicon diodes can be connected (D_5 and D_6 in Fig. 8) mit v_i to about 0·5 volt. If gold-bonded germanium diodes used this voltage can be limited to 0·2 volt, thus making all les of the switch non-conducting; silicon diodes do not duct in either direction at 0·2 volt.

o close the diode switch a positive voltage is applied to the inal x_1, and a negative voltage of the same magnitude to inal x_2 of Fig. 9(*a*). This switching voltage can be many s smaller than the voltage to be sampled, v, and need only arge enough to cause conduction, provided that R_p is kept iciently large and R_d sufficiently small to ensure negligible rs. This is the case because a positive input voltage v_i er than the positive switching voltage will block D_3 but D_4. Hence the switching voltage appears at the input of amplifier as a positive voltage v', which after amplification inversion reduces the voltage v_i and all diodes become con- ting. A similar reasoning shows that the switch will close n the voltage v_i is negative and larger than the switching age. It should be noted that if the limiting diodes D_5 and are used their effect is negligible when the switch is closed, e v_i almost equals v' and both are very small. For example, an amplifier gain, K, of 100 000 and an output of 100 volts, voltage v_i would be 1 mV and silicon diodes have very high stances for such small voltages of both polarities; germanium des, however, may increase the output due to drift in the plifier because of their relatively low resistances. The tages which operate the diode switches are shown in . 10, and the overall accuracy of the sampler is many times

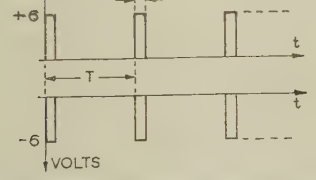


Fig. 10.—Switching voltages required to operate a diode switch.

better than 0·1% provided that the operational amplifiers are drift-corrected.

Before concluding the description of the sampler, it is worth while to mention that a single silicon transistor can perform the function of the four-diode switch, with the additional advantages of lower switching power and single-pulse switching. The circuit is as shown in Fig. 11, where the diodes D_5 and D_6 are

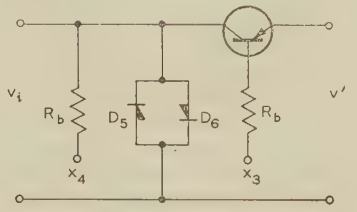


Fig. 11.—Transistor switch.

again used to limit the input voltage to the switch, v_i, to 0·5 volt. The transistor switch is closed when a negative voltage larger than 0·5 volt is applied to terminal x_3. However, the switching voltage causes an amplifier output determined by the base resistance R_b and the feedback resistances R_f and R_h of the operational amplifier, and this output can be eliminated by applying a fixed positive voltage of equal magnitude to terminal x_4. To open the transistor switch a positive voltage larger than 0·5 volt must be applied to terminal x_3, but care should be taken that the blocking resistance between the base and the emitter is sufficiently large effectively to block this voltage. The computer described in the paper employs only silicon diode switches, and transistor switches were not considered because silicon transistors were not available when the work was commenced; the base-emitter blocking resistance of germanium transistors is not sufficiently large, and large errors may result owing to temperature variations.

(3.2) Sampler Hold

Holding is equivalent to demodulation or restoration of the continuous signal and normally precedes the plant to be controlled and always follows the discrete controller (Fig. 4). The circuit of the sampler hold (Fig. 12) is similar to that of the sampler and the same diode gate [Fig. 9(*a*)] is used for switching.

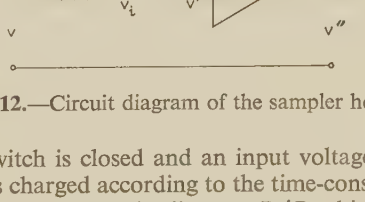


Fig. 12.—Circuit diagram of the sampler hold.

When the switch is closed and an input voltage v is applied, the capacitor is charged according to the time-constant C_fR_f and the final steady-state output is $v'' = -vR_f/R_i$; this applies for an ideal switch and very large amplifier gain. Considering the equivalent circuit of the diode switch [Fig. 9(*b*)] it can be readily shown that again the effect of the parallel resistance R_p on the steady-state value is negligible if the amplifier gain is sufficiently large. However, a low value of R_p may considerably increase the output due to drift in the operational amplifier. The main effect of R_d is only a slight increase in the charging time-constant.

Since a time-constant is involved, the sampling interval δ must be sufficiently large to allow the final value to be reached. The feedback resistance cannot be too small because the power

output of the amplifier is limited, and the minimum feedback capacitance is dependent upon the sampling period T and the required accuracy as will be shown later. This imposes a lower limitation on the sampling period δ for a given sampling period T, but ratios of $\delta/T = 1/1\,000$ can be obtained if required.

When the switch opens, the output of the amplifier remains at the sampled value for a considerable length of time. Capacitor leakage, amplifier drift, grid current, amplifier input impedance and finite amplifier gain will tend to change the amplifier output. For example, in a sampler-hold with an amplifier gain K of 100 000, a feedback capacitance of $0{\cdot}1\,\mu$F, an amplifier input resistance of 10^9 ohms and if the maximum output is 100 volts, the holding time must be limited to 10 000 sec for an accuracy of 0·1%. In addition, a capacitor leakage time-constant of 10^5 sec, or an amplifier drift of 10 mV or a grid current of 10^{-10} amp, will limit the holding time to 100 sec for the same accuracy and the above values of gain, capacitance, input resistance and maximum output. A larger capacitance will proportionally increase the holding time; e.g. an $1\,\mu$F capacitor with a time-constant of 10^6 sec will permit holding times up to 1 000 sec under the same conditions. However, sampling periods as long as this are seldom encountered in control systems; they are normally not longer than a few seconds. The use of a drift-corrected amplifier removes drift effects, and if a capacitor input is employed grid-current effects are also eliminated.

The change in output due to a voltage v_i at the input of the diode switch and the leakage resistances of the silicon diodes D_2 and D_4 is small for short sampling periods. If long holding times are required, this voltage can always be reduced to a negligible value by the use of the limiting diodes, as pointed out in Section 3.1.

(3.3) Unit Time Delay

It was already shown that a pulse transfer function can be implemented by means of summing amplifiers and unit time delays, i.e. time delays equal to one sampling period. The sampler hold described in the previous Section has the properties of a time delay, since the signal at the kth sampling instant is available at the output of the amplifier just prior to the $(k + 1)$th sampling instant, but unfortunately not exactly at this instant. This difficulty can be avoided by connecting two sampler-hold circuits SH_1 and SH_2 in series (Fig. 13), each one being sampled at equal but time-shifted sampling periods. The switching pulses

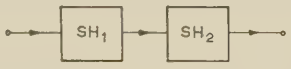


Fig. 13.—Unit time delay made up of two sampler-hold circuits in series.

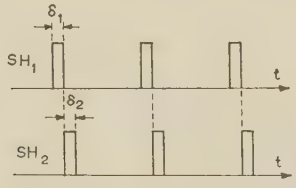


Fig. 14.—Time sequence of the switching pulses of a unit time delay.

for SH_1 and SH_2 are shown in Fig. 14, from which it is clearly seen that the output of SH_2 at the $(k + 1)$th sampling instant is equal to the input to SH_1 at the kth sampling instant. This is a time delay equal to one sampling period, and it is noted that there is no net sign inversion from the input to the output of the unit time delay, since each operational amplifier and thus each sampler hold introduces a negative sign. The sampling interval δ_2 of the second sampler hold, can be times larger than the sampling interval δ_1 of the first sa hold, and this permits the realization of very small sam ratios δ_1/T for a given accuracy of a unit time delay. pulsed analogue computer was made up of such time d and summing operational amplifiers.

(4) ANALOGUE COMPUTER

(4.1) Description

The analogue computer was built at the University of South Wales as a research project to permit studies i behaviour of sampled-data systems. The computer can be to determine the response of pulse transfer functions, to sim sampled-data systems and operate as a discrete controll compensation of feedback control systems.

A total of 20 operational amplifiers are employed, eig which are used to provide four unit time delays; one is us sampler, one as sampler hold and the remaining amplifie be used as summing amplifiers or to simulate the plant computing components, i.e. the input and the feedback ele which determine the function of an operational amplifie mounted on plug-in units and can be easily modified wh different mode of operation is required. Provision is ma balancing, monitoring, initiation and other similar fun which are well-known analogue computer techniques are not be discussed here. A central patching board is availab interconnection of the operational amplifiers.

All switching is performed with four-diode switches empl unselected silicon diodes. The long-term drift of such sw is only a few millivolts, and its effect is negligible for an o accuracy of 0·1% of a computing operation. The swit voltage supply, which provides the required positive and neg pulses to close the diode switches during the sampling int was built entirely of transistors (Fig. 15).

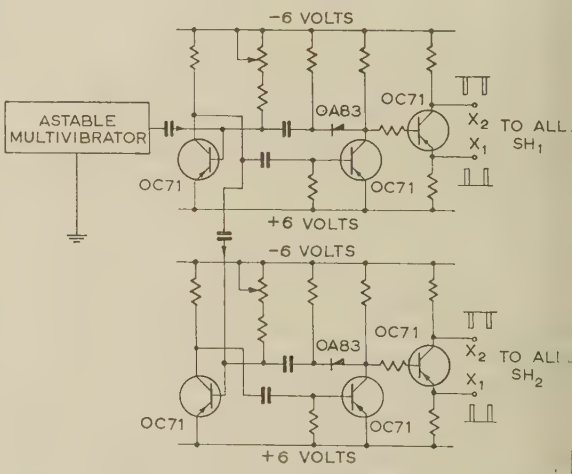


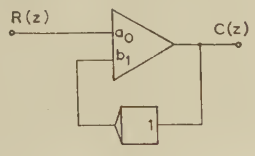
Fig. 15.—Circuit diagram of the switching voltage supply

Two monostable multivibrators each driving an output consisting of a single transistor are employed. When the vibrators are at rest, the output transistors are fully condu and they are so balanced that their output terminals approximately +50 mV and −50 mV. For such low vo the silicon diodes are non-conducting in both directions a switches are effectively open. When the multivibrato triggered the output transistors become non-conductin output terminals are now at +6 and −6 volts and the switches are closed. Each multivibrator is adjustable, s

pulse width, i.e. the sampling interval δ, can be varied from 20 millisec.

ne sampling period depends entirely upon the frequency of triggering pulses, but there is a limit at higher frequencies the sampling period cannot be smaller than the smallest pling interval, which, in this case, is 3 millisec. There is no at lower frequencies. To make the computer independent ternal oscillators an astable multivibrator was added, whose ency can be varied between 1 and 50 c/s.

ne astable multivibrator only triggers the monostable multi-



16.—Computer arrangement of a pulse transfer function with a single negative real pole.

ator, whose output operates the diode switches of the first pler hold (Fig. 13) of a unit time delay. This multivibrator operates the switches of the separate sampler and the rate sampler hold in the computer. The switch of the nd sampler hold SH_2 of a unit time delay is being operated ne other monostable multivibrator. This multivibrator must riggered after the switch of the first sampler hold SH_1 has ned; this is accomplished by using the pulse which opens the

(4.2) Operation

To illustrate the operation of the computer in determining the response of pulse transfer functions, consider the simple transfer function $D(z) = a_0/(1 + b_1 z^{-1})$, where a_0 and b_1 are real and positive. The computer arrangement of this transfer function is shown in Fig. 16, and it is seen that the correct sign for b_1 is available at the input of the summing amplifier since the unit time delay introduces no net phase shift. A record of the step response of this computer arrangement when followed by a sampler hold is shown in Fig. 17 for various values of b_1; it should be noted that the overshoot of the recorder pen is absent when the output is observed on an oscilloscope. It is seen that, for $b_1 > 1$, the output grows in an oscillatory manner indicating that the computer arrangement represents an unstable system. In such a computer arrangement, even the slightest disturbance will cause an output which increases until limited by saturation. Making use of the unit time delay as an adder, a zero can be sometimes introduced in the transfer function without the use of an additional amplifier. From Fig. 18 the output of this computer arrangement is

$$C(z) = -a_0 R(z) - a' b_1 z^{-1} R(z) - b_1 z^{-1} C(z),$$

which yields

$$-\frac{C(z)}{R(z)} = \frac{a_0 + a' b_1 z^{-1}}{1 + b_1 z^{-1}}$$

and the zero is

$$\left(-\frac{a' b_1}{a_0}\right)$$

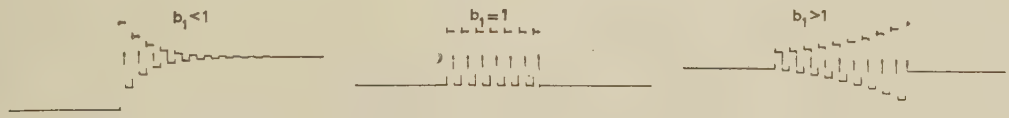


Fig. 17.—Step response of the computer arrangement in Fig. 16, for various values of gain b_1.

ch of SH_1 to trigger the multivibrator, so that the switch of closes immediately after the switch of SH_1 has opened . 14).

may be noted that the impedance of the output stages is zero, and currents fed into the switching voltage supply, to the voltages v_i at the inputs of the diode switches, may rb the balance. Again, this effect can be eliminated by the of the limiting diodes D_5 and D_6, as shown in Fig. 8. Another ibility is to design the voltage supply so that blocking ages of equal magnitude are applied to the diodes to open switches. This will eliminate currents fed back into the ching voltage supply, but the blocking voltages must be at t equal to the voltages to be sampled and more elaborate age supply circuits will be required. In addition, the large king voltages may cause diode currents far larger than the current of an amplifier, and since the blocking resistances ne diodes D_2 and D_4 are, in practice, never equal, considerable rs may result, as indicated in Section 3.1.

he overall accuracy of the computer is largely a matter of bration, but it is better than 1% if standard high-stability stors are employed as computing components; this, of course, lies only if the accuracy requirements discussed in the vious Section have been met. When an accuracy of 0·1% is ired, careful adjustment of the various computing gains is essary, and if this accuracy is to remain over long periods of e, drift-free operational amplifiers may have to be employed. wever, many industrial controls do not require an accuracy er than 1%, and in research and simulation this accuracy ften quite satisfactory also.

Transfer functions with a single positive real pole can also be realized with the computer arrangement of Fig. 16, but with an inverting amplifier in the feedback loop. Another arrange-

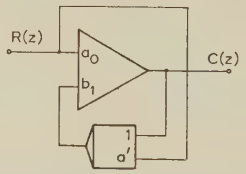


Fig. 18.—Computer arrangement of a pulse transfer function with a negative real pole and a negative real zero.

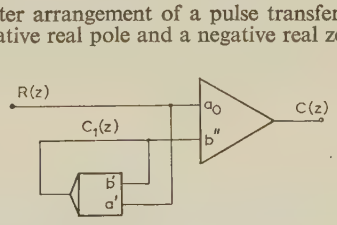


Fig. 19.—Computer arrangement of a pulse transfer function with a positive real pole and a real zero.

ment which eliminates the inverting amplifier is shown in Fig. 19. The sampler or sampler hold, which always follows the arrangement of a pulse transfer function, can be used as a summing amplifier, so that, in fact, such transfer functions are realized with a single unit time delay. From Fig. 19,

$$C_1(z) = a' z^{-1} R(z) + b' z^{-1} C_1(z) \quad \text{or} \quad C_1(z) = \frac{a' z^{-1} R(z)}{1 - b' z^{-1}}$$

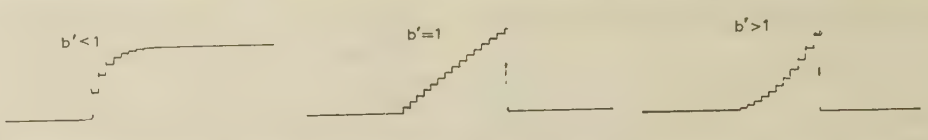


Fig. 20.—Step response of the computer arrangement in Fig. 19, for $a'b'' = b'a_0$ and various values of gain b'.

and $C(z) = -b''C_1(z) - a_0R(z)$ gives the transfer function as

$$-\frac{C(z)}{R(z)} = \frac{a_0 + (a'b'' - b'a_0)z^{-1}}{1 - b'z^{-1}}$$

It is clear that, in addition to the desired pole b', a zero can be introduced on the real axis, by choosing the gains a' and b'' so that $-(a'b'' - b'a_0)/a_0$ is equal to this zero. When $a'b'' = b'a_0$ the transfer function is $a_0/(1 - b'z^{-1})$, and the step response of this computer arrangement for various values of b' is shown in Fig. 20. Again, the solution for $b' > 1$ is unstable, and it is noticed that when $b' = 1$ the computer arrangement represents an integrator, since the output at the sampling instants is the same as the pulse sequence of the integral of the step input.

Two unit time delays are needed to realize a pair of complex conjugate poles. Transfer functions with large numbers of poles can be realized by the general method indicated in Section 2, but often the partial-fraction method is preferred because the effect of the individual poles can be studied independently.

(5) DESIGN OF DISCRETE CONTROLLER TO COMPENSATE AND AVOID SATURATION IN A FEEDBACK SYSTEM

There are a number of methods available for the design of a discrete controller to compensate a linear feedback system.[1,2,3,4] In saturating systems the controller can be designed in such a manner that the system does not saturate when subjected to the most severe input expected; for smaller inputs the system is obviously linear. This approach to the design of a discrete controller has been discussed by Bergen and Ragazzini[2,3] and Mullin.[8] The method of design presented below generally follows the procedure developed by Mullin, but it is simpler to use, and serves to illustrate the compensation of a feedback system by the analogue computer.

The system to be compensated with a discrete controller contains a saturating element as shown in Fig. 21, and for a step input an overall response is desired such that the output is equal to the input for all time after a finite number of sampling periods q.

The input to the plant consists of discrete steps, i.e. the input is constant between sampling instants. Hence the output of the plant $c(t)$ can be obtained as the sum of responses due to pulses of height $H(k)$, giving the infinite summation

$$c(t) = \sum_{k=0}^{q-1} H(k)h(t - kT) + \sum_{k=q}^{\infty} H(k)h(t - kT)$$

where $h(t)$ is the response of the plant due to a unit pulse input, i.e. an input pulse of unit height and width equal to one sampling period T.

If the plant contains at least one integration, the second term is zero, and the first term can be constrained so that it yields equations which meet the two conditions of the desired overall response. The steady-state condition requires that the sum of all time-independent terms is equal to the input, and the transient condition requires that the sum of all time-dependent terms is equal to zero after the specified number of sampling periods. From these equations a pulse sequence of the forcing function $H(z)$ can be determined, with pulses of magnitude exceeding its saturation limits. Next the pulse sequence of error function $E(z)$ is obtained as the difference of the input the system $R(z)$ and the output of the plant $C(z)$ due to

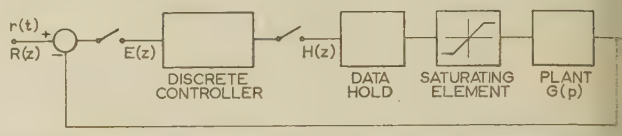


Fig. 21.—Feedback system with a discrete controller and a satur element.

above determined forcing function $H(z)$. Finally, the transfer function of the discrete controller is given by

$$D(z) = \frac{H(z)}{E(z)}$$

Example 1.—The plant transfer function is

$$G(p) = \frac{1}{p(p + 1)}$$

and saturation occurs whenever the pulses of the forcing fun $H(z)$ exceed ± 1 unit. It is desired to compensate the syste a maximum step input of 2 units, and after four sampling pe of 1 sec the output is to equal the input for all time; the re for the choice of four sampling periods for the given inp discussed at the end of this example. The step response o



Fig. 22.—Step response of the sampled uncompensated linear feed system in Example 1.

sampled but uncompensated linear system is shown in Fig and clearly the oscillatory behaviour of the system is undesi

The response of the plant to a unit step input is

$$w(t) = t - 1 + \varepsilon^{-t}$$

and the response of the plant to a unit pulse input become

$$h(t) = w(t) - w(t - T)$$
$$= t - 1 + \varepsilon^{-t} - [(t - T) - 1 + \varepsilon^{-(t-T)}]$$

Hence the output of the plant due to an input pulse seq with pulses of magnitude $H(k)$ is given by

$$c(t) = \sum_{k=0}^{3} H(k)[(t - kT) - 1 + \varepsilon^{-(t-kT)} - (t - T - kT) + 1 - \varepsilon^{-(t-T-k}$$

or, after cancellation of terms, by

$$c(t) = \sum_{k=0}^{3} H(k)[T + \varepsilon^{-t}(\varepsilon^{kT} - \varepsilon^{T+kT})]$$

the output is to equal the input after four sampling periods, or $c(t) = 2{\cdot}00$ after $t = 4T$, the steady-state condition res that

$$c_{ss}(t) = \sum_{k=0}^{3} H(k)T = 2{\cdot}00$$

or the transient to die out,

$$\varepsilon^{-t} \sum_{k=0}^{3} H(k)(\varepsilon^{kT} - \varepsilon^{T+kT}) = 0$$

ere are two equations and four unknowns in the pulse nce of the forcing function. It follows that two pulses be chosen arbitrarily with the understanding, of course, hey never exceed the saturation limit of ± 1 unit. Assuming the first pulse $H(0)$ and the second pulse $H(1)$ take the ng value of $+1$ unit, the equations become

$$\sum_{=0}^{3} H(k)T = 1 + 1 + H(2) + H(3) = 2{\cdot}00$$

$$\sum_{=0}^{3} H(k)(\varepsilon^{kT} - \varepsilon^{T+kT}) = (\varepsilon^{0} - \varepsilon^{1}) + (\varepsilon^{1} - \varepsilon^{2}) + H(2)(\varepsilon^{2} - \varepsilon^{3}) + H(3)(\varepsilon^{3} - \varepsilon^{4}) = 0$$

lving the equations for the two unknowns yields $= 0{\cdot}29$ and $H(3) = -0{\cdot}29$, and the pulse sequence of forcing function becomes $H(z) = 1{\cdot}00 + 1{\cdot}00z^{-1} + z^{-2} - 0{\cdot}29z^{-3}$.

determine the pulse sequence of the plant output $C(z)$ o the forcing function $H(z)$, the pulse sequence of the unit response is found by calculating the values of $w(t) = 1 + \varepsilon^{-t}$ at $t = kT$, giving

$$= (0 + 0{\cdot}37z^{-1} + 1{\cdot}14z^{-2} + 2{\cdot}05z^{-3} + 3{\cdot}02z^{-4} + \ldots)$$

pulse sequence of the plant response due to a unit pulse t is then

$$z) = w(z) - w(z)z^{-1} = (0{\cdot}37z^{-1} + 0{\cdot}77z^{-2} + 0{\cdot}91z^{-3} + 0{\cdot}97z^{-4} + \ldots)$$

the pulse sequence of the plant output becomes

$$z) = H(0)h(z) + H(1)h(z)z^{-1} + H(2)h(z)z^{-2} + \ldots$$

$$= (0{\cdot}37z^{-1} + 1{\cdot}14z^{-2} + 1{\cdot}79z^{-3} + 2{\cdot}00z^{-4} + 2{\cdot}00z^{-5} + \ldots)$$

he error function is determined from $E(z) = R(z) - C(z)$, for a step input

$$R(z) = 2{\cdot}00(1{\cdot}00 + 1{\cdot}00z^{-1} + 1{\cdot}00z^{-2} + \ldots)$$

the above $C(z)$, it is found to be

$$E(z) = 2{\cdot}00 + 1{\cdot}63z^{-1} + 0{\cdot}86z^{-2} + 0{\cdot}21z^{-3}$$

he discrete controller transfer function is then

$$D(z) = \frac{H(z)}{E(z)} = \frac{1{\cdot}00 + 1{\cdot}00z^{-1} + 0{\cdot}29z^{-2} - 0{\cdot}29z^{-3}}{2{\cdot}00 + 1{\cdot}63z^{-1} + 0{\cdot}86z^{-2} + 0{\cdot}21z^{-3}}$$

he analogue-computer arrangement of the controller is wn in Fig. 23, and a record of the step response of the compensated system is shown in Fig. 24, which should be compared the step response of the uncompensated system in Fig. 22. here is a maximum input for which the system can be compensated in four sampling periods without exceeding the specified

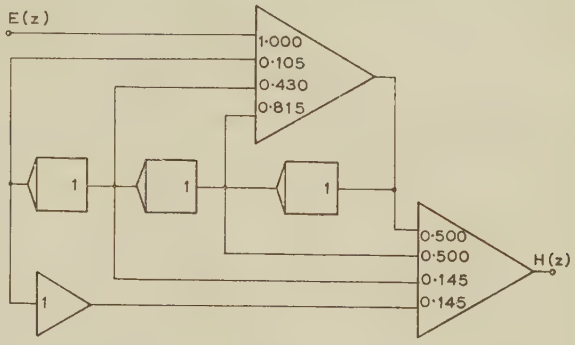


Fig. 23.—Analogue computer arrangement of the discrete controller in Example 1.

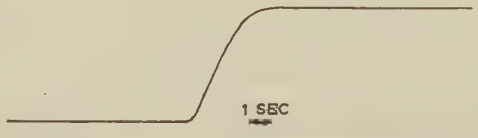


Fig. 24.—Step response of the compensated system in Example 1.

limit of ± 1 unit for the forcing function. To find this maximum input R_{max}, the steady-state equation is written as

$$1 + 1 + H(2) + H(3) = R_{max}$$

but the transient equation remains as before. Elimination of $H(3)$ yields $R_{max} = 1{\cdot}815 + 0{\cdot}636H(2)$. The input will be a maximum when $H(2)$ is positive and has the maximum possible value, i.e. when $H(2) = 1{\cdot}00$, giving

$$R_{max} = 1{\cdot}815 + 0{\cdot}636 = 2{\cdot}451$$

The value of the last pulse is then readily determined as $H(3) = -0{\cdot}549$. Had the solution resulted in a value of $H(3) > |1|$, the procedure would have had to be repeated by eliminating $H(2)$ and determining R_{max} for the maximum possible value of $H(3)$. A similar procedure is adopted when a larger number of unknown pulses are present.

If the maximum input for which it is desired to compensate the system is larger than R_{max}, the number of sampling periods must be increased, and if the maximum input is much smaller than R_{max}, a decrease in the number of sampling periods may be possible, resulting in improved performance of the system.

Example 2.—This example illustrates the design method when the plant transfer function has complex poles. The plant transfer function is

$$G(p) = \frac{p^2 + 2{\cdot}5p + 1}{p(p^2 + 2p + 5)}$$

and again the controller must be designed so that the forcing function never exceeds ± 1 unit for a step input. It is required that the output shall equal the input after five sampling periods. The sampling period T is, as before, 1 sec.

The response of the plant to a unit step input is

$$w(t) = 0{\cdot}20t + 0{\cdot}42 + 0{\cdot}46\varepsilon^{-t} \sin(2t - 1{\cdot}15)$$

and the response of the plant to a unit pulse input is given by

$$h(t) = 0{\cdot}20t + 0{\cdot}42 + 0{\cdot}46\varepsilon^{-t} \sin(2t - 1{\cdot}15) - 0{\cdot}20(t - T) - 0{\cdot}42 - 0{\cdot}46\varepsilon^{-(t-T)} \sin[2(t - T) - 1{\cdot}15]$$

The output of the plant due to a pulse sequence $H(k)$ is

$$c(t) = \sum_{k=0}^{4} H(k)h(t - kT)$$

and the steady-state and the transient conditions yield three equations:

$$\sum_{k=0}^{4} H(k)0{\cdot}20T = R_{max}$$

$$\sum_{k=0}^{4} H(k)[\varepsilon^{kT}\cos 2kT - \varepsilon^{T+kT}\cos 2(T + kT)] = 0$$

$$\sum_{k=0}^{4} H(k)[\varepsilon^{kT}\sin 2kT - \varepsilon^{T+kT}\sin 2(T + kT)] = 0$$

Hence three pulses $H(2)$, $H(3)$ and $H(4)$ of the forcing function are unknown, with $H(0) = H(1) = 1{\cdot}00$. Substitution of numerical values and elimination yields $R_{max} = 0{\cdot}68$, with $H(2) = 1{\cdot}00$, $H(3) = 0{\cdot}31$ and $H(4) = 0{\cdot}093$. The discrete controller can now be designed as before.

In conclusion, it may be said that the analogue computer confirmed the validity of the design method. This method can also be adapted to design a controller for ramp inputs with zero steady-state error when the plant to be controlled has only one integration. Similarly, a controller can be designed for step inputs with zero steady-state error when the plant has no integration. In these cases the additional integration required for zero steady-state error is contained in the discrete controller. The response of a compensated system to inputs larger than the maximum input for which the system has been designed can be determined by a step-by-step procedure, but it is tedious and subject to accumulative errors. The analogue computer can readily provide the solutions to such inputs, because a saturating element can be easily introduced in a computer arrangement.

(6) CONCLUSIONS

It was shown that analogue computer techniques can be employed to solve pulse transfer functions or compensate control systems in a discrete manner. The operation is purely electr and only operational amplifiers are used. With the incre reliability of operational amplifiers by the use of transisto is possible to employ such computers for discrete compens of industrial controls. In the study and research of sampled- systems the analogue computer provides better insight int behaviour of a system than digital computers normally pe with the additional advantages of low cost and simplicit programming.

(7) ACKNOWLEDGMENT

The author wishes to thank Professor R. E. Vowels of School of Electrical Engineering, University of New S Wales, who initiated the project, for permission to publish paper.

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DISCUSSION BEFORE THE MEASUREMENT AND CONTROL SECTION, 15TH NOVEMBER, 196

Dr. J. H. Westcott: The devices S1 and S2 in Fig. 1, although looking like switches, do not act as such. For mathematical convenience in the analysis, we require that the output of the sampler should be a pulsation train, the pulse areas being equal to the sampled ordinates.

In the error-sampled feedback control system with discrete controller, shown in Fig. 4, it would have been helpful if the supply-disturbance as well as the load-disturbance responses had been considered, since both are of importance in process control.

The author's comment that 'In general, the use of a digital computer to compensate a single system would be uneconomical' was true of servo mechanisms, but no longer of process plant.

In Section 2 the author states: 'The necessary time delay in the above computer arrangement, which is equivalent to storage in digital computers, has been the main obstacle to the development of discrete analogue computers'. This is not quite true. The cup-and-bucket delay-line idea has been considered for ten years, and certainly used for a number of years. One takes a capacitor, charges it, and it hands on its charge to its neighbour along the line. This is what the author is doing, except that he is using only a single cup-and-bucket pair at a time.

Dr. R. H. Barker: There is certainly no lack of applications for this subject. In the electricity supply industry, for example, we already have automatic control of small units, and t being extended to major portions of a station and perha due course to major portions of a system. Much of this inv sampled, and incidentally, telemetered data.

In tracing the history of sampled data-control systems would probably start with Hazen, who, in 1934, pointe some of the advantages of applying definite correctio intervals. The now well-known z-transform was introduc Hurwicz, who during the war was concerned with the pr of making the best use of the inherently intermittent p radar information. Next came the digital servo syste applications such as machine-tool control. The main adva is that the accuracy can be very high, for it is limited o the accuracy of the position encoder and not by the contr Position encoders have been developed to read to at 15 binary digits or 1 in 32 000.

However, there are many cases where 0·1% accura adequate, and we are asked to consider the introducti sampling, not because it is inherent in the data source, an because it is imposed by a digital computer, but in its own We have indeed come full circle.

It is therefore worth while to examine the advantages type of control provided by the system of Fig. 4 relative

ct (i.e. continuous) feedback system with the usual shaping vork to provide the desired performance characteristics. erficially, it would appear that quite a lot of information is arded in the sampling process, and that the performance of sampled system should therefore be inferior to the best tinuous system. This would only be true, of course, if the ortant time-constants were all much less than the sampling od, and it is a consequence of the well-known theorem ting sampling frequency to bandwidth.

'an it be shown, however, that sampling enables a positively erior performance to be achieved? I do not believe that adequate answer has ever been given to this question. There he fact that the transient resulting from an input such as a t step or unit ramp can be halted after a limited number of pling periods, whereas the response of the continuous system away gradually. However, this may not be an important erion of performance, and I believe this whole question uld be thoroughly examined.

ince the unit time delay is at the heart of the method of trol it is a great pity that so much equipment—two operaal amplifiers with their attendent electronic switches—are uired to represent it. The cause would be greatly advanced omeone could suggest a simple and elegant alternative.

have myself used a device which is accurate to rather better n 1%, and is based on the humble cathode-follower (see . A).

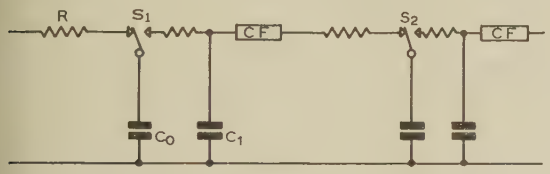


Fig. A

Two units of delay are shown. S_1 and S_2 operate simulteously, changing position for a brief period every sampling tant. C_0 is charged through R during the period prior to tching. The charge on C_0 is the measure of the information, d this is shared with C_1. The charge on C_1 is held for the uing data period while C_0 is being charged to some fresh ue. The cathode-follower provides a low-impedance source drive the next delay stage.

The capacitance C_1 must, of course, be much smaller than t of C_0, say one-tenth of it, and ideally C_1 should be disurged each time. If this is not done, the pulse transfer function the stage is not $1/z$ but

$$\frac{1}{z-\delta}$$

ere $$\delta = \frac{C_1}{C_0 + C_1}$$

is, in general, possible to allow for this in the way that the tputs from the delay stages are utilized. Thus, if this type of ay unit were adopted in the system of Fig. 18, the pole would ur at

$$z = -(b_1 - \delta)$$

d the zero at $$z = -\left(\frac{a'b_1}{a_0} - \delta\right)$$

e effect of δ can thus be exactly compensated by adjustment the other constants of the system. Alternatively, if a pole required at $z = +\delta$, where $\delta < 1$, it is only necessary to select correct ratio of C_1/C_0.

With regard to Section 5, I can see little point in designing a system to just reach saturation with any input that is less than the maximum that can be catered for. Hence the only case of interest in Example 1 is where R_{max} is introduced. In this case the plant is driven as hard as possible during the first three data periods of the step function, and just brought to rest during the fourth. Presumably the object is to achieve the final value as quickly as possible, and I wonder whether some further improvement would not be possible by, for example, optimizing the proportion of the total transient response time allowed for the initial positive part of the forcing function.

Mr. B. M. Brown: I would suggest an alternative approach to that given in Section 5 which not only simplifies the computation but gives a clearer idea of the underlying principles. This makes use of z transforms.

If $G^*(z)$ is the pulse transfer function of the combination of the data hold and the plant,

$$E(z) = R(z) - G^*(z)H(z)$$

and $$D(z) = \frac{H(z)}{E(z)}$$

For a system with a finite settling time, $E(z)$ and $H(z)$ must be polynomials in z^{-1}.

Considering Example 1 of the paper,

$$G(p) = \frac{1}{p(p+1)}$$

so that $$G^*(z) = \frac{1}{z-1} - \frac{1-e^{-1}}{z-e^{-1}}$$

If $r(t)$ is a step function of amplitude r_0, $R(z) = \frac{r_0 z}{z-1}$.

Let $H(z) = h_0 + h_1 z^{-1} + h_2 z^{-2} + h_3 z^{-3}$

$$= z^{-3}(h_0 z^3 + h_1 z^2 + h_2 z + h_3)$$

and $$E(z) = z^{-3}(e_0 z^3 + e_1 z^2 + e_2 z + e_3)$$

Then $$z^3 E(z) = \frac{r_0 z^4 - z^3 H(z)}{z-1} + \frac{(1-e^{-1})z^3 H(z)}{z-e^{-1}}$$

But the left-hand side is a polynomial, and for the right-hand side to be a polynomial, $z-1$ must be a factor of $r_0 z^4 - z^3 H(z)$, and $z - e^{-1}$ a factor of $z^3 H(z)$. Then, by the remainder theorem, $H(1) = r_0$, and $e^{-3}H(e^{-1}) = 0$,

i.e. $$h_0 + h_1 + h_2 + h_3 = r_0 \quad . \quad . \quad . \quad . \quad \text{(A)}$$

$$e^{-3}h_0 + e^{-2}h_1 + e^{-1}h_2 + h_3 = 0 \quad . \quad . \quad . \quad \text{(B)}$$

If these conditions are satisfied $E(z)$ can be found directly by dividing out the two partial fractions. Thus, if $r_0 = 2$, $h_0 = h_1 = 1$ and $h_2 = -h_3 = 0{\cdot}29$, the working is as follows using synthetic division:

	2	−1	−1	−0·29	0·29		1	1	0·29	−0·29
(×1)		2	1	0	0·29	(×0·37)		0·37	0·51	0·29
	2	1	0	−0·29	0		1	1·37	·80	0

We have therefore

$$z^3 E(z) = (2z^3 + z^2 - 0{\cdot}29) + 0{\cdot}63(z^2 + 1{\cdot}37z + 0{\cdot}80)$$

or $$E(z) = 2 + 1{\cdot}63z^{-1} + 0{\cdot}86z^{-2} + 0{\cdot}21z^{-3}$$

The discrete controller obtained in this way will give zero transient at all sampling instants after the fourth. Since, after this instant, the input to the plant is constant, the output must

be of the form $A + Be^{-t}$. There can therefore be no inter-sample ripple.

The problem of finding suitable values of h_0, h_1, h_2 and h_3, taking account of the limitations imposed by saturation, can be solved using a topological method.

Since these four quantities satisfy the two conditions (A) and (B), two of them can be chosen arbitrarily. Let these be h_0 and h_1. Then h_2 and h_3 are given by

$$h_2 = \frac{r_0}{1 - e^{-1}} - h_0(1 + e^{-1} + e^{-2}) - h_1(1 + e^{-1})$$

$$h_3 = \frac{-e^{-1}r_0}{1 - e^{-1}} + e^{-1}(1 + e^{-1})h_0 + e^{-1}h_1$$

If $|h_0|$, $|h_1|$, $|h_2|$, $|h_3| \leqslant 1$, possible points in the (h_0, h_1) plane must lie between each of the following pairs of lines: $h_0 = \pm 1$, $h_1 = \pm 1$,

$$h_0(1 + e^{-1} + e^{-2}) + h_1(1 + e^{-1}) = \frac{r_0}{1 - e^{-1}} \mp 1 \quad . \quad \text{(C)}$$

$$h_0(1 + e^{-1}) + h_1 = \frac{r_0}{1 - e^{-1}} \pm e \quad . \quad . \quad . \quad \text{(D)}$$

With $r_0 = 2$, (C) and (D) become

$$\begin{aligned} 1{\cdot}50h_0 + 1{\cdot}37h_1 &= 2{\cdot}18 \text{ or } 4{\cdot}18 \\ 1{\cdot}37h_0 + h_1 &= 5{\cdot}90 \text{ or } 0{\cdot}46 \end{aligned}$$

The only possible values of h_0 and h_1 are therefore those inside the triangle PQR (see Fig. B).

The effect of varying r_0 is to change the positions of (C) and (D), without changing their directions. Clearly the maximum value of r_0 for which settling can be achieved in four limited steps is obtained when the line $h_2 = 1$ passes through P.

Dr. A. J. O. Cruickshank: With discrete compensation, quite apart from the very real practical advantages of simplicity and low cost of pulsed analogue computers compared with pure digital computers, there is the very great advantage of being able to formulate comparatively simple methods of synthesis and design of control systems. Not the least of these is the possibility of remaining wholly in the time domain. I believe that time-domain analysis has not received sufficient attention from analytical experimenters, and I welcome a return to such analysis.

One very important advantage in using a pulsed analogue computer as the physical realization of a required discrete compensator is that the required coefficients of the polynomials in

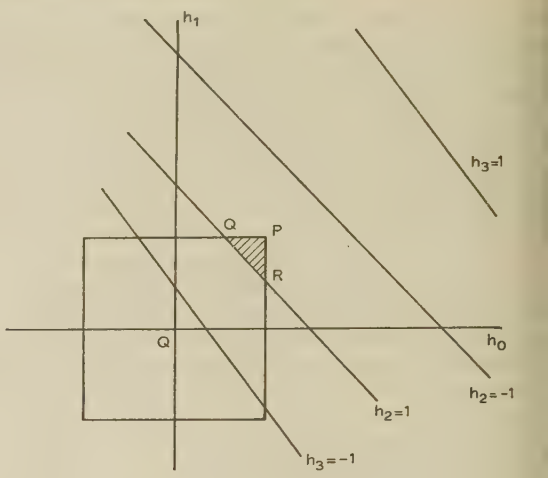


Fig. B

z^{-1} appear explicitly as the gains of the various channels. is in direct contrast to Sklansky's pulsed *RC*-networks met and makes for a very easy understanding of design.

Much of the design work with discrete compensators has concentrated on aperiodic test inputs, such as the step, r etc. The attempts which have been made to optimize a de with a statistical description of the input have returned more to the frequency domain, power spectra, etc., which to bring more than usual complexity to sampled-data syste

I should like to suggest, following to some extent the re work of Prasad,* that it is possible to remain wholly in the domain when optimizing a system with statistical inputs. fact, the results of such an analysis should give specifically ordinates of the pulse response of the desired compensato say, the first ten intervals following zero, after which the nates may be neglected. By definition, these ordinates des the pulse-transfer function, up to the tenth power of z^{-} the required compensator. Furthermore this can be rea immediately by a sequence of unit delays and gains, as indic in the paper. We should also remember that sampled- techniques may be used to design continuous systems.

I therefore see an immediate future for pulsed analogu unit delay computers, not only in furthering simply and che the practical realization of sampled-data systems, but als tools for trying obvious analytical and arithmetic procedure

THE AUTHOR'S REPLY TO THE ABOVE DISCUSSION

Mr. T. Glucharoff (*in reply*): I am interested to learn that sampled-data controls are being extended to major portions of a system. The sampled-data theory is well developed and sampled controls are being gradually accepted in the industry, but I fully agree with Dr. Barker that a thorough investigation of the performance criteria of sampled-data systems as compared with continuous systems is required. There are, however, other advantages with sampled systems, some of which have been pointed out by Dr. Cruickshank, and another is the possibility of time sharing of equipment, as with digital computers for discrete compensation.

In reply to Dr. Westcott's remark regarding the devices S1 and S2 in Fig. 1, it appears that the requirement that the pulse areas should be equal to the sampled ordinates has not been stressed adequately in the paper. Mathematically, the pulses are represented by impulses of area equal to the sampled ordinates, and in practical samplers it is required that the sampling interval δ should be short compared with the time variatio the sampled signal and the time-constants of a following tinuous system. Under these conditions S1 and S2 ca represented by switches if the sampling interval δ is 1 sec practice, all samplers are some form of switch, and the value of the sampling interval is of little or no importance a sampler is followed by a holding circuit, which is usuall case. If a sampler is followed by a continuous system large time-constant, the gain of the sampler is approximate

The method employed to realize the required unit delays in the analogue computer is essentially the cup-bucket method, as pointed out by Dr. Westcott. Use has made of this method by means of operational amplifiers electronic switches, so that an accuracy of better than w 0·1% has been obtained. If the operational amplifiers

* PRASAD, T.: 'Analysis and Optimisation of a Class of Non-Linear St Systems for Random Processes', Proceedings of First Congress of IFAC, M 1960 (to be published).

aced by cathode-followers in open-loop operation, sampling, ling and time delay could have been obtained in a manner lar to that described in the paper, but the accuracy of the vidual operations would have been about 1%. The use of rational amplifiers permits closed-loop operation and lers the circuit independent of the characteristics of the tronic switches, requires smaller switching voltages, pro- s low output impedance, and the gain becomes solely de- dent upon the input and the output resistances. If the olifiers are drift-corrected, long-term operation is possible grid-current effects can be eliminated. It should also be ed that a sampler or a sampler hold is very frequently used ne same time as adder, and it was shown in the paper that, by cing use of a unit time delay as an adder, the total number operational amplifiers in some computer arrangements can be reduced. However, this is certainly no reason to assume that a simpler time delay of comparable accuracy is not required.

The alternative design method given by Mr. Brown makes use of the z-transform and is a valuable approach to the design of discrete controllers in feedback systems. The topological method of determining suitable values for the magnitudes of the pulses of the forcing function, however, may become difficult if the number of arbitrary pulses is not limited to two.

The speakers have suggested that it would be desirable to determine the effects of supply and load disturbances, to consider optimizing the transient response and to apply time-domain statistical design theory to the design of discrete controllers. This clearly shows that there is no lack of ground for further work in the field of sampled-data systems.

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THE STANDARDIZATION OF INTERNATIONAL MICROWAVE RADIO-RELAY SYSTEMS

By W. J. BRAY, M.Sc.(Eng.), Member.

(*The paper was first received* 21*st July, and in revised form* 11*th October*, 1960.)

SUMMARY

Microwave radio-relay systems using line-of-sight paths now form a substantial part of the international trunk network, both for multi-channel telephony and for television. In order that such systems shall provide satisfactory transmission over long distances they must conform with certain minimum performance standards. Furthermore, in order to facilitate the interconnection of radio-relay systems with one another and with line systems, certain common characteristics are necessary. The establishment of links across national frontiers also requires the use of equipment with similar characteristics in each country.

The paper outlines the work of the International Radio Consultative Committee (C.C.I.R.) and the International Telegraph and Telephone Consultative Committee (C.C.I.T.T.) in defining preferred characteristics for microwave radio-relay systems using line-of-sight paths, and discusses the technical and other reasons for the particular values adopted. The need for preferred characteristics for radio-relay systems using tropospheric-scatter propagation is also referred to.

(1) INTRODUCTION

(1.1) Types of Microwave Radio-Relay System

The microwave radio-relay systems forming the subject of the paper operate at frequencies above about 1 Gc/s, and are of two types:

(*a*) Systems using line-of-sight, or nearly line-of-sight, paths.[1–6]*
(*b*) Systems employing the tropospheric-scatter mode of propagation with paths extending well beyond the visible horizon.[7–8]*

Radio-relay systems using line-of-sight paths with distances of up to about 80 km between repeater stations are already well established in many countries as part of the national trunk networks for multi-channel telephony and television. When interconnected with similar systems or with line systems in other countries they become part of the international trunk network. Such radio-relay systems may provide on each radio carrier up to 2 000 telephone channels or a television channel with or without some hundreds of telephone channels, and a number of radio carriers may be provided over a given route. Large-capacity radio-relay systems are used mainly in countries with highly-developed communication systems; however, low-capacity systems are often used in countries with smaller communication requirements and as spurs to large-capacity systems. Thus there is a need for line-of-sight radio-relay systems covering a wide range of capacities. The international standardization of such systems is well advanced and is the main subject of the paper.

Radio-relay systems using the tropospheric-scatter mode of propagation have so far found application only on a limited scale. Nevertheless, they are useful for the provision of links up to some 500 km long in a single hop over stretches of water or difficult terrain, under conditions where line-of-sight ra relay links or coaxial-cable links are impracticable or un nomic. Capacities of up to 100 or so telephone channels television channel of moderate quality can be provided on radio carrier on links up to about 300 km long. Bec tropospheric-scatter systems are a comparatively recent dev ment, international standardization has not yet reache advanced state, but there is urgent need for this to be d for reasons to be discussed later.

Future developments, including radio-relay systems with l capacity than is provided on systems at present in use modulation methods offering greater spectrum economy discussed in Section 9.

(1.2) Parts Played by the C.C.I.R. and the C.C.I.T.T. i Standardization

The International Radio Consultative Committee (C.C.I. of the International Telecommunication Union (I.T.U.) played an important part in the standardization of internat radio-relay systems, this work being entrusted to Study Grou of the C.C.I.R. In addition, the International Telegraph Telephone Consultative Committee (C.C.I.T.T.) has mad possible to secure a large measure of international agreemen common overall performance objectives for radio-relay and systems, so that both may be integrated on an equal footing i the general trunk network. Where necessary, special joint w ing or study groups comprising both line and radio system exp have been set up to study certain problems. One such grou the C.C.I.T.T./C.C.I.R. Joint Working Party on Circuit No which has been appointed to determine the permissible level circuit noise in radio-relay and line systems for the transmiss of telephone and telegraph signals. Another joint study gro known as the Joint Study Group for the Transmission of T vision (C.M.T.T.), has been formed to study the problems the transmission of television signals by line and radio-re systems.

The problems involved in standardization are studied firs national level in the various countries by groups of exp representing the administration, industry and the opera agencies (including the broadcasting organizations) concer with telecommunication. These groups, which are o organized by the administration concerned on the same bas the C.C.I.R. Study Groups, prepare proposals for considera in the Study Groups of the C.C.I.R. and subsequent consid tion at the Plenary Meetings of the C.C.I.R. which are about once every three years. In many cases interim St Group meetings are held during the period between Ple Meetings in order to prepare for the latter. By a proces discussion, detailed examination of the technical factors invo and compromise where this is justified, it has been found poss to secure international agreement, formulated as the 'Rec mendations' of the C.C.I.R.,* on the more important cha teristics of line-of-sight radio-relay systems, and a start has b

* General references are given in Section 11.1.

* References to the Recommendations and Reports of the C.C.I.R. are giv Section 11.2.

This is an 'integrating' paper. Members are invited to submit papers in this category, giving the full perspective of the developments leading to the present practice in a particular part of one of the branches of electrical science.
Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.

Mr. Bray is in the Post Office Engineering Department.

e on the characteristics of the newer systems such as those
lving tropospheric-scatter propagation.

should be noted that the Recommendations of the C.C.I.R
not mandatory; in other words, the technical characteristics
mmended by them are 'preferred' rather than obligatory.
example, it is open to any two administrations wishing to
blish a radio-relay link across a common frontier to do so
g either C.C.I.R. preferred characteristics or other charac-
tics, by mutual agreement. In practice, the preferred
acteristics are widely used, not only for international links
ning frontiers but also for national systems within the
tiers of individual countries, because the planning, provision
maintenance of links are thereby facilitated.

his paper contains a summary of the preferred characteristics

(iii) Intermediate-frequency characteristics.
(iv) Modulation characteristics of the r.f. carriers.
(v) Radio-frequency characteristics, including the frequency patterns of the channels in multi-r.f.-channel systems.
(vi) Characteristics of the supervisory, control and monitoring systems needed for maintenance purposes.

The significance of these characteristics can best be seen from a consideration of the general arrangement of a typical radio-relay system accommodating several r.f. channels.

(1.4) General Arrangement of Typical Multi-R.F.-Channel Radio-Relay System

The general arrangement of a typical multi-r.f.-channel radio-relay system shown in Fig. 1 and described below is applicable

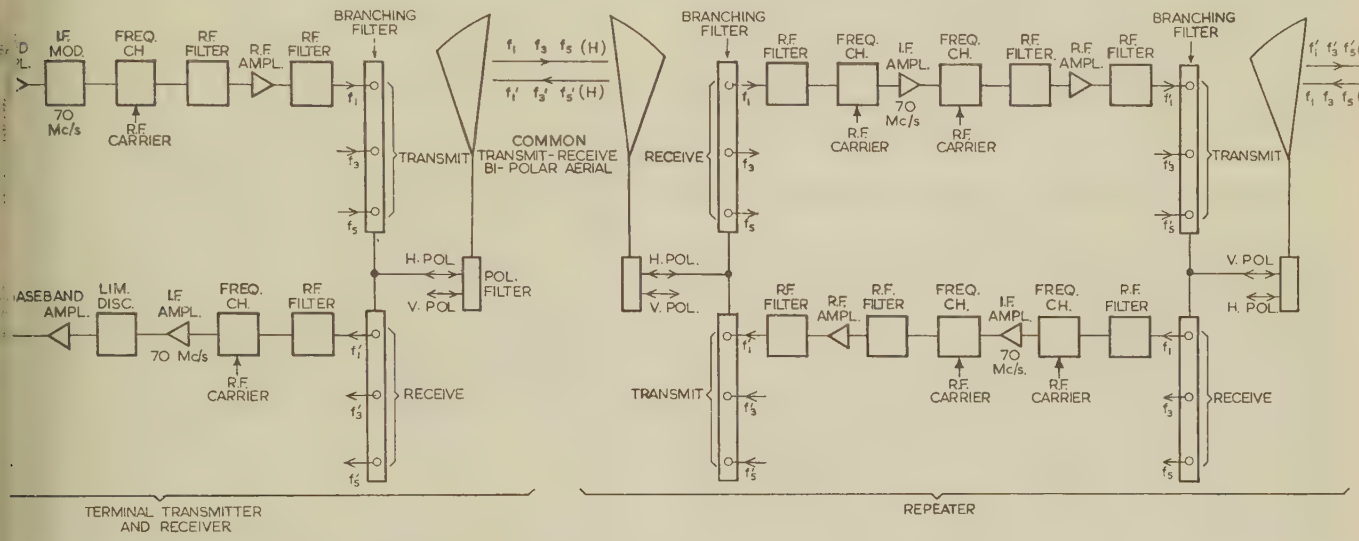


Fig. 1.—Schematic arrangement of typical multi-r.f.-channel radio-relay system.

ommended by the C.C.I.R. and discusses the technical and erational reasons why they have been adopted. Since the commendations of the C.C.I.R. can, for reasons of space, be ven only in summarized form, reference should be made to the ginal documents, Volumes I, II and III of the IXth Plenary sembly of the C.C.I.R. (Los Angeles, 1959) and to the documents of the C.C.I.T.T., since these are the definitive texts.

It is important to remember that all transmission systems are bject to a process of evolution, and radio-relay systems are exception. Revision of certain of the preferred characteristics ll undoubtedly be required in the light of new technical velopments and as demands on the available frequency space cessitate still greater economy in the use of the spectrum.

(1.3) Characteristics to be Agreed on an International Basis

Certain basic characteristics of radio-relay systems need to be reed on an international basis to facilitate the establishment international links and to permit the interconnection of radio-lay systems with one another and with line systems (Report o. 131).* These characteristics are as follows:

(i) Overall transmission performance for telephony (including music circuits), television and telegraphy (including data transmission).
(ii) Baseband characteristics.†

* The Recommendations and Reports referred to thus are given in Volumes I d III of the Documents of the IXth Plenary Assembly of the C.C.I.R. (Los Angeles, 59). They are available at present in English and French.

† The 'baseband' of a radio-relay system may be defined as the frequency band the input and output of the system which is occupied by the multi-channel telephony, levision or telegraphy signals and any pilots or other signals transmitted by the stem.

primarily to line-of-sight systems, but certain of the features will also apply to other systems, e.g. those involving tropospheric-scatter propagation.

Most of the line-of-sight microwave radio-relay systems used for international circuits employ frequency modulation of the r.f. and i.f. carriers.[3, 9] The reasons why frequency modulation is widely used in preference to other methods of modulation are as follows:

(*a*) The signal/noise ratio in the telephony or television channels is greater than can be obtained using amplitude modulation under comparable conditions.
(*b*) A linear relation between the input and output amplitudes in the i.f. and r.f. stages is not essential with frequency modulation whereas it is with amplitude modulation.
(*c*) Limiting in receivers and repeaters can be used to prevent changes in the amplitude of the baseband output signal, even under conditions of fast and deep fading of the r.f. carrier.
(*d*) The bandwidth required is less than would be needed by pulse-modulation systems of comparable capacity and performance.

Where several baseband signals, each comprising a block of f.d.m. telephony channels or a television channel, are to be transmitted over the same route, considerable savings in cost can be achieved by using the same aerials and feeders to transmit and receive a number of r.f. carriers on different frequencies, as shown in Fig. 1. It is often convenient first to generate the f.m. carrier at an intermediate frequency, generally centred on 70 Mc/s, and then to shift the frequency to the desired radio frequency by mixing with an r.f. carrier of good frequency stability. After selection in r.f. filters and amplification to a

power of up to a few watts, the r.f. signals from a number of transmit channels can be combined in a branching filter and applied via a common feeder to a wideband aerial. The same wideband aerial may, if desired, function simultaneously for reception and transmission. On the receiving side the r.f. signals are separated by r.f. filters, changed to an intermediate frequency generally centred on 70 Mc/s, and then amplified to a suitable level for limiting. The limited signal is applied to a linear discriminator, the output of which is the desired baseband signal. Other facilities which may be provided, but which are not indicated in Fig. 1, comprise automatic frequency control of the transmitted and received carrier frequencies and automatic gain control prior to limiting.

The schematic arrangement of a typical multi-r.f.-channel repeater of the non-demodulating type is also shown in Fig. 1. That shown involves frequency-changing from a radio frequency to an intermediate frequency, also generally centred on 70 Mc/s, and back to a different radio frequency. Other forms of repeater are sometimes used in which the change of radio frequency is effected directly without the use of an intermediate frequency.

(1.5) Economic and Operational Aspects of Multi-R.F.-Channel Systems

In considering possible alternative arrangements of multi-r.f.-channel radio-relay systems from the standpoint of international standardization, certain economic and operational aspects must be taken into account.

From the economic point of view it is desirable to keep the total equipment provision to a minimum by using, whenever possible, common equipment shared between several broadband channels. On the other hand, too much common equipment may reduce the overall reliability below an acceptable standard or give rise to design difficulties leading to inferior performance.

The achievement of adequate reliability generally necessitates a protection broadband channel which can be manually or automatically switched into service if a working broadband channel fails, e.g. due to propagation or equipment failures. Since the introduction of a protection channel greatly reduces the out-of-service time, it follows that a number of working channels, e.g. up to five, can share the same protection channel and still achieve adequate reliability.

Cost savings may be achieved by accommodating several r.f. channels on the same aerial, and by using the same aerials for transmission and reception. Thus a satisfactory r.f. channel plan (see Section 5) should make provision for common-aerial working.

(1.6) Interconnection of Radio-Relay and Line Systems

(1.6.1) Multi-Channel Telephony Systems.

Most line systems for multi-channel telephony use frequency-division multiplexing (f.d.m.), the signal in each telephone channel being of single-sideband suppressed-carrier form. The channel spacing is 4 kc/s, and the basis of combining telephone channels in groups (12 channels), supergroups (60 channels) and mastergroups (300 channels) has been standardized by the C.C.I.T.T. (originally the C.C.I.F.); typical arrangements of f.d.m. telephony channels are shown in Fig. 2.

A great deal of line plant in the form of 12-, 24-, 48- and 60-circuit carrier systems and 600- and 960-circuit coaxial-cable systems using the standard C.C.I.T.T. f.d.m. arrangements of channels has been brought into use by telephone administrations in many countries during the past three decades. It has been natural, therefore, to develop multi-channel radio-relay systems on the same f.d.m. channelling basis as for line systems, since this permits the interconnection of line and radio-relay systems

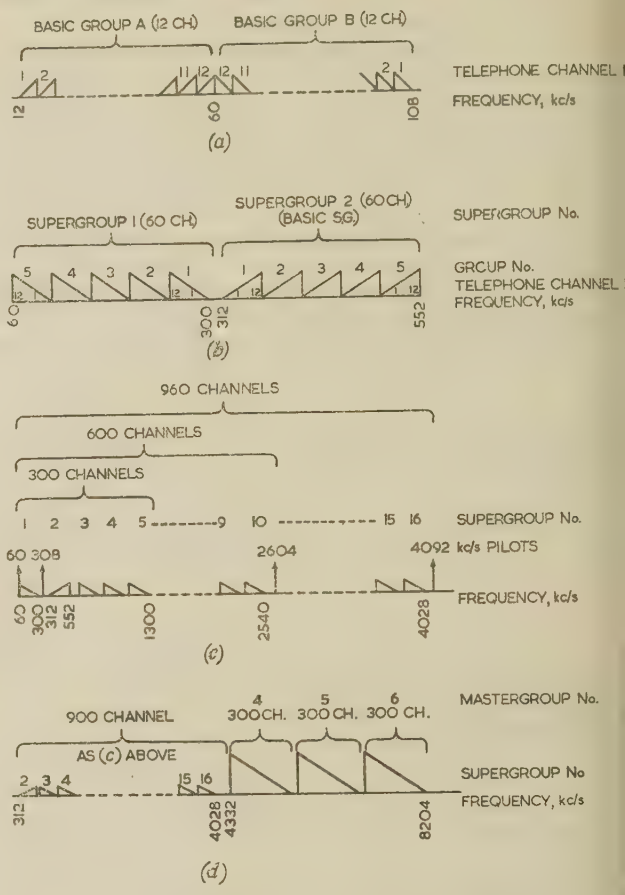


Fig. 2.—Typical arrangements of f.d.m. telephony channels.

(*a*) 12- and 24-channel systems.
(*b*) 60- and 120-channel systems.
(*c*) 300-, 600- and 960-channel systems.
(*d*) 1 800-channel system.

at the group, supergroup or mastergroup levels, without need to translate channels to audio frequency. It also per blocks of circuits to be readily switched, e.g. in the even breakdown, from line to radio-relay systems or vice versa, enables the same type of translating equipment to be used both types of system.

For the interconnection of radio-relay and line systems desirable to choose suitable f.d.m. signal levels and ci impedances at the output and input of the radio-relay syst The levels recommended by the C.C.I.R. and C.C.I.T.T. (tion 3.1) provide relatively low input levels and relatively output levels in order to facilitate interconnection via stan group and supergroup translating equipment. This arrange provides considerable flexibility since it enables groups or su groups to be dropped or inserted at the point of interconnec Alternatively the low input level can be used, if desired accommodate the losses in tail-cable connections, e.g. betwe radio-relay-link terminal station and a line repeater stati few miles away, as shown, for example, in Fig. 3.

(1.6.2) Television Systems.

The long-distance transmission of television signals coaxial-cable systems generally requires the translation of signal from video frequency to a frequency band above a hundred kilocycles per second in order to avoid low-frequ interference and to simplify repeater design. The signal t

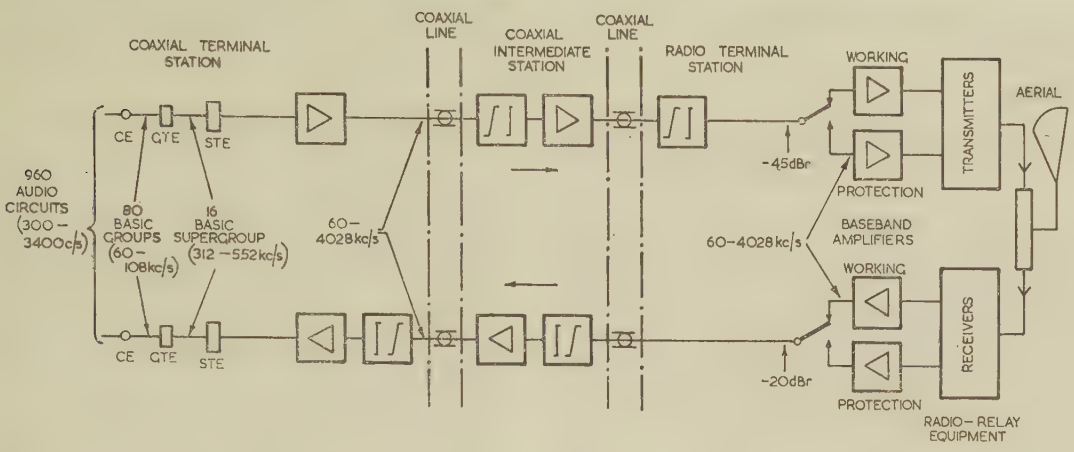


Fig. 3.—Interconnection of coaxial and radio-relay systems for 960 telephony circuits.

STE = Supergroup translating equipment.
GTE = Group translating equipment.
CE = Channel equipment.

In the example shown all 960 circuits are translated to audio frequency. In practice, groups, supergroups or mastergroups may be abstracted for routing as required. Pilots, gain-regulating amplifiers, etc., are not shown.

ted by the cable consists of an asymmetric-sideband signal h a video carrier which may be of reduced level. On the er hand, radio-relay systems transmit television signals most ciently when the video signal, with or without pre-emphasis, dulates the radio carrier directly. Modulation of the radio rier by the video signal directly provides a better signal/noise io than is possible if an asymmetric-sideband signal modulates carrier, and avoids possible difficulties due to harmonics of video carrier generated by residual non-linearity in the radio-ay system.

Thus interconnection between radio-relay and coaxial-cable tems is usually best carried out at video frequency, although mmetric-sideband carrier signals may be transmitted by dio-relay systems of adequate linearity and signal/noise ratio. Typical television interconnection arrangements between axial-cable and radio-relay systems are shown in Fig. 4.

To facilitate interconnection between radio-relay systems and e systems, preferred video levels and circuit impedances at e point of interconnection, and the appropriate transmission ndwidths for television signals with various line standards, ve been defined (see Section 2.3).

(2) TRANSMISSION PERFORMANCE OF INTERNATIONAL MICROWAVE RADIO-RELAY SYSTEMS

(2.1) Multi-Channel Telephony Systems

The transmission performance of the longest international circuits used for telephony should be such that, taking into account any additional limitations imposed by the national circuits to which the international circuits are connected, almost all subscribers are able to communicate without difficulty. It is not practicable or economic to ensure that *all* possible pairs of subscribers would have no difficulty in talking over the longest connections likely to be used in practice, but nevertheless the aim is to ensure that unsatisfactory calls constitute a negligible percentage of the total.

The overall (audio-to-audio) transmission characteristics which should be provided on international telephone circuits have been defined by the C.C.I.F. (now the C.C.I.T.T.).[10] For present purposes the relevant characteristics are as follows:

(i) Audio band effectively transmitted.
(ii) Non-linear distortion.
(iii) Phase distortion (variation of the group delay across the band effectively transmitted).

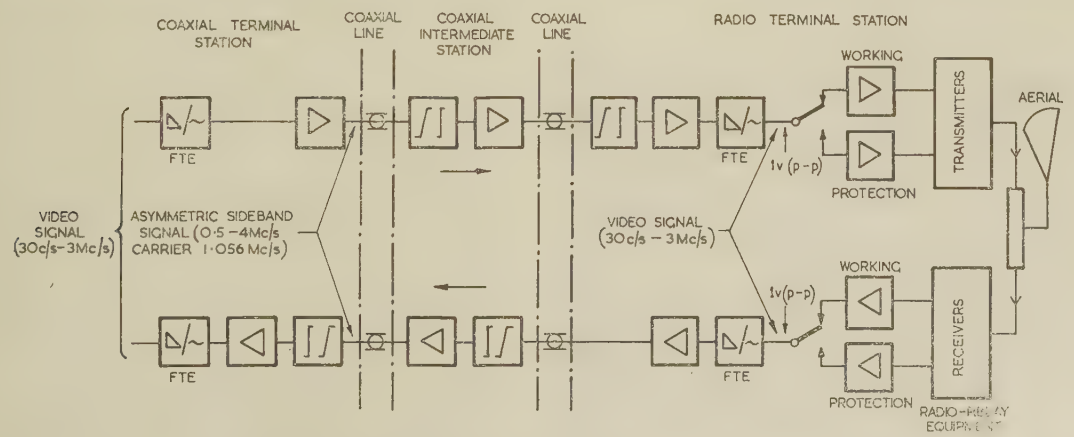


Fig. 4.—Interconnection of coaxial and radio-relay systems for television (405-line system).

FTE = Frequency translating equipment.
Pilots, gain regulating amplifiers and pre-emphasis/de-emphasis networks are not shown.

(iv) Audio group delay (propagation time).
(v) Echoes.
(vi) Gain stability.
(vii) Crosstalk.
(viii) Circuit noise.

These overall audio characteristics apply irrespective of the means of transmission, whether by line, coaxial cable or radio relay.

(2.1.1) Audio Bandwidth, Non-Linear Distortion and Phase Distortion.

In the case of microwave radio-relay systems, the audio bandwidth, 300–3 400 c/s, and the phase distortion are determined primarily by the f.d.m. terminal equipment and are not limited by the radio-relay system itself, since the latter is intrinsically a broadband transmission medium. Furthermore, when the crosstalk requirements discussed in Section 2.1.5 are met this usually ensures that the audio non-linear distortion is sufficiently small.

(2.1.2) Group Delay.

The effective velocity of transmission in a microwave radio-relay system approaches that of light (3×10^8 m/s) since most of the transmission path is in free space; the present provisional limit of 100 millisec for the group delay on an international circuit would thus permit the use of microwave radio-relay links up to some 25 000 km in length (this allows a few milliseconds for delays in repeaters, translating equipment, audio tail cables, etc.).

(2.1.3) Echoes.

Since completely separate 'go' and 'return' transmission paths operating on different frequencies are normally used on microwave radio-relay systems, any audio echoes usually occur outside the radio-relay systems, e.g. due to hybrid unbalance at the audio 4/2-wire terminations. Echoes due to crosstalk are effectively excluded if the crosstalk requirements discussed in Section 2.1.5 are met.

(2.1.4) Gain Stability.

The C.C.I.F. has indicated that the overall stability of the chain of national and international circuits, measured between the terminal trunk exchanges, should provisionally be not worse than 0·2 neper (1·74 dB). Although overall gain control using group or supergroup pilots is commonly used on such systems, high baseband-to-baseband gain stability is nevertheless required on individual radio-relay systems. The specification of the gain-stability requirements for individual systems is at present being studied by the C.C.I.T.T.

(2.1.5) Crosstalk.

Crosstalk, i.e. the appearance in an audio channel of unwanted signals from another audio channel, may be either linear or non-linear. Linear crosstalk is due in general to unwanted inductive, capacitive or resistive couplings between circuits operating at the same frequency. Linear crosstalk may occur between the audio circuits themselves, in the frequency-translating terminal equipment or between the basebands of radio-relay systems. It rarely occurs in the intermediate or radio-frequency equipment of radio-relay systems. The near- or far-end linear crosstalk ratio regarded as acceptable by the C.C.I.F. in the case of international cable circuits is 6·7 nepers (58·2 dB) for 90% of such circuits, and 6 nepers (52·1 dB) for all circuits; these values are also applicable to circuits on international microwave radio-relay systems. Non-linear crosstalk may occur as a result of a non-linear relation between the baseband input and output amplitudes of a radio-relay system; it results in unwanted signals in a given channel in the baseband (and thence in the corresponding audio channel) due to harmonics and intermodulation products of signals in other channels in the same baseband. If there are many interfering channels the resulting crosstalk is si to random noise, and it is convenient to discuss the permi level of non-linear crosstalk with that of circuit noise Section 2.1.6).

(2.1.6) Circuit Noise and Non-Linear Crosstalk.

The C.C.I.F. has indicated that it is desirable that on national circuits the total circuit noise (including non-l crosstalk) should not exceed a psophometric e.m.f. of 2 m a point of relative level −0·8 neper (−7 dB).[11] This c sponds approximately to a mean psophometric powe 10 000 pW at a point of zero relative level. The interpret of this recommendation in terms of radio-relay systems an use of hypothetical reference circuits for such systems is cussed in Section 2.2.

(2.2) Hypothetical Reference Circuits and Circuit Noise F.D.M. Multi-Channel-Telephony Radio-Relay System

The permissible level of the total circuit noise (including linear crosstalk) is the most important single factor in the d of a radio-relay system. For a system of given capac determines the power of the transmitters, the size of the ae the maximum path length that can be used and thus the c the system. It is highly desirable, therefore, that the spec tion for total circuit noise should not be more stringent th necessary for the intended use, in this case for the transmi of telephony (including signalling tones).

It has long been the practice of the C.C.I.F. to specif design objectives for noise on international cable circu terms of a hypothetical reference circuit (h.r.c.) as an aid t designers and planners of such systems, and this has continued by the C.C.I.T.T. The h.r.c. is divided into a nu of equal-length sections, with various specified stages of tra tion to audio, basic group and supergroup interposed be the sections. This practice has been applied by the C.C to radio-relay systems. The hypothetical reference circuit (*a*) f.d.m. radio-relay systems with capacities of 12 to 120 phone channels per r.f. carrier, and (*b*) f.d.m. radio- systems with capacities of more than 120 telephone channel r.f. carrier, are shown in Fig. 5, each circuit being 2 500 km

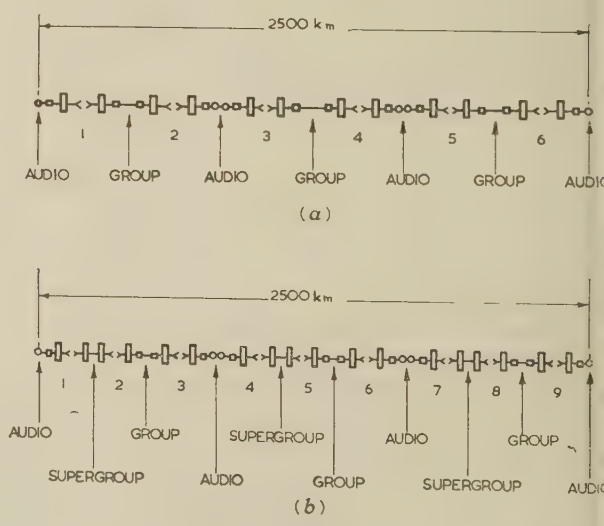


Fig. 5.—Hypothetical reference circuits for f.d.m. multi-chann telephony radio-relay systems.

(*a*) 12–120 telephone channel capacity per r.f. channel.
(*b*) More than 120 telephone channel capacity per r.f. channel.

—○— Channel modulator or demodulator.
—□— Group modulator or demodulator.
—▯— Supergroup modulator o demodulator.
——< Radio modulator or demo dulator with baseban input or output.

divided into (*a*) six homogeneous sections and (*b*) nine mogeneous sections (Recommendations Nos. 285, 286 in ion 11.2).

adapting to radio-relay systems the noise design objectives ified by the C.C.I.T.T. for cable systems, the C.C.I.R. has n into account the following factors (Recommendation 287):

(*a*) The circuit noise on radio-relay systems comprises two mponents: (i) basic noise arising mainly in the input circuits receivers and (ii) intermodulation noise arising mainly from osstalk between the telephony channels due to non-nearity.[13, 14]

(*b*) The basic noise at the output of a radio-relay system aries with time due to fading on the various sections of the stem, and it must therefore be expressed statistically and referably in relation to a relatively long period of time such a month.

(*c*) The intermodulation noise depends upon the loading of e system by the f.d.m. telephony signal, and it therefore aries with time, being greatest during the busy hour. It ust also be expressed statistically.

(*d*) The f.d.m. signal during the busy hour can be satisctorily represented by a uniform spectrum (white-noise) gnal, which, at a point of zero relative level, has a level of:

$(-15 + 10 \log_{10} N)$ dBm for 240 channels or more,
$(-1 + 4 \log_{10} N)$ dBm for 12 to 240 channels,*

here N is the number of traffic channels (see Fig. 15).

he C.C.I.R. has recommended that the noise power at a nt of zero relative level in any telephone channel on a 0 km hypothetical reference circuit for f.d.m. radio-relay ems should not exceed the provisional values given below commendation No. 287):

(*a*) 7 500 pW psophometrically weighted mean value in any our.†

(*b*) 7 500 pW psophometrically weighted 1 min mean power or more than 20% of a month when the fading is severe.

(*c*) 47 500 pW psophometrically weighted 1 min mean power or more than 0·1% of a month when the fading is severe.

(*d*) 1 000 000 pW unweighted (with an integrating time of millisec) for more than 0·01% of a month when the fading s severe.

he value of 7 500 pW given above is derived from the 00 pW noise allowed in the complete international circuit ction 2.1.6) by deducting 2 500 pW for the noise appropriate the f.d.m. translating equipment. The '1 min mean' time-stant is of the same order as the average duration of telephone s and the 'integrating time of 5 millisec' takes account of the racteristics of signalling and telegraphy systems.

n order to apply the above recommendation to real radio-ty system, as distinct from the h.r.c., the following additional ommendations have been made (Recommendation No. 288):

a) The psophometrically weighted noise power at a point of o relative level in the telephone channels of a real f.d.m. io-relay system of length L, where L is between 250 and 0 km, and whose composition does not differ appreciably m the hypothetical reference circuit, should not:

(i) exceed $3L$ picowatts mean value in any hour.
(ii) exceed $3L$ picowatts 1 min mean power for more than 20% f any month.

b) As a planning objective, and recognizing that the permance achieved for very short periods of time by real circuits when installed may differ from the planning objective and is difficult to measure precisely, the psophometrically-weighted noise power at a point of zero relative level in the telephone channels of such a real system as is considered above should not exceed 47 500 pW 1 min-mean power for more than $(L/2\,500) \times 0{\cdot}1\%$ of any month.

It will be noted that the mean values of the noise power appropriate to an hour, or more than 20% of a month, are subdivided according to the length of the system involved, whereas in the case of the relatively high noise levels exceeded only for small percentages of time it is the latter which is subdivided according to the length of the system. Subdivision of the time is considered justified on the grounds that deep fades producing high levels of noise are generally of such short duration that they occur only for very small percentages of time and are unlikely to coincide (Report No. 130). These simple rules give a reasonably satisfactory basis for the design of long radio-relay systems. A more detailed and precise basis of design applicable to short as well as long systems has been given elsewhere.[15]

The foregoing recommendations are design objectives for radio-relay systems with line-of-sight paths. Recommendations relating to other systems, namely those using tropospheric-scatter propagation, are referred to in Section 8.6.

(2.3) Television Systems

The requirements for the transmission of television signals over long distances have been embodied in a recommendation prepared by the C.M.T.T. and adopted by the C.C.I.R. (Recommendation No. 267). The main points in this recommendation which are of primary interest to the designers and planners of radio-relay systems are summarized below.

The requirements are specified in terms of a long-distance international circuit (BC in Fig. 6), which comprises a chain of

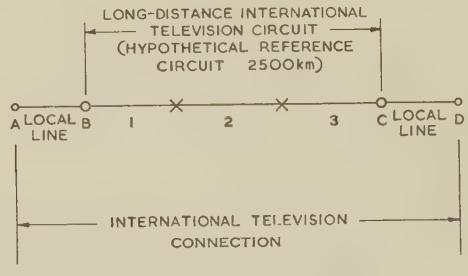


Fig. 6.—International television connection and hypothetical reference circuit for television.
○ Terminal video points.
× Intermediate video points.

national and international television cable or radio links. As a basis for the planning and design of systems a 2 500 km hypothetical circuit has been defined which comprises three equal-length sections, with video interconnections.

(2.3.1) Requirements at Video Interconnection Points.

At video interconnection points the input and output impedance is to be 75 ohms resistive (nominal value), unbalanced to earth, with a return loss of at least 24 dB relative to 75 ohms. The polarity of the signal at such points is to be positive, i.e. such that black-to-white transitions are positive-going. The nominal peak-to-peak amplitude of the video signal (including the synchronizing signal) is 1 volt, the nominal amplitude of the synchronizing signal being 0·3 volt (Recommendation No. 270).

(2.3.2) Transmission Performance of the Hypothetical Reference Circuit.

The following transmission requirements apply to the complete

Provisional value for systems with less than 60 channels.
The level of uniform-spectrum noise power in a 3·1 kc/s band must be reduced 2·5 dB to obtain the psophometrically-weighted noise power.

hypothetical reference circuit. The basis for subdivision in terms of the requirements for partial circuits is at present being studied by the C.C.I.T.T.

(2.3.2.1) *Gain and Stability.*

A long-distance international circuit having the form of the hypothetical reference circuit is required to have, at the time of setting up, an insertion gain of 0 ± 1 dB. The variation of insertion gain with time should not exceed $\pm 0{\cdot}3$ dB for short-period (e.g. 1 sec) variations, and $\pm 1{\cdot}0$ dB for medium-period (e.g. 1-hour) variations.

(2.3.2.2) *Signal/Random-Noise Ratio.*

The required minimum signal/noise ratio, X, for continuous random noise, expressed as the ratio in decibels of the peak-to-peak amplitude of the picture signal (excluding the synchronizing signal) to the r.m.s. amplitude of the noise within the range between 10 kc/s and the nominal upper frequency limit of the system, f_c, is given in Table 1. The values given assume the use of a specified low-pass filter and weighting network and a measuring instrument having an 'effective time-constant' or 'integrating time', in terms of power, of 1 sec. The values shown are minima to be exceeded for all except 1% of a month.

Table 1

VIDEO BANDWIDTH AND SIGNAL/RANDOM-NOISE RATIO FOR VARIOUS TELEVISION SYSTEMS

	Number of lines	405	525	625	625	819	819
System	Nominal upper limit of video frequency band f_c (Mc/s)	3	4	5	6	5	10
Signal/weighted-noise ratio X, dB ..		50	—	52	57	—	50

Signal/noise ratios for 525-line and 819-line (5 Mc/s bandwidth) systems have not yet been defined by the C.C.I.R.

The interpretation of this recommendation in terms of radio-relay systems is discussed in Section 2.4.

(2.3.2.3) *Signal/Periodic-Noise Ratio.*

In the case of periodic noise the required minimum signal/noise ratio is specified as the ratio in decibels of the peak-to-peak amplitude of the picture signal (excluding the synchronizing pulses) to the peak-to-peak amplitude of the noise. For power-supply hum this ratio should exceed 30 dB; for single-frequency noise between 1 kc/s and 1 Mc/s the ratio should exceed 50 dB, falling linearly to 30 dB at f_c (25 dB at f_c in the case of 405-line systems). Reference should be made to Recommendation No. 267 for the values appropriate to the 819-line (10 Mc/s bandwidth) system.

(2.3.2.4) *Non-Linearity Distortion.*

'Line-time' non-linearity distortion of the picture signal is measured with a test signal in which the 'picture' portion of every fourth line consists of a sinusoidal signal of frequency $0{\cdot}2f_c$ on a sawtooth wave, the intermediate lines being set to either white or black level. The magnitude of the distortion is indicated by the ratio of the minimum to the maximum peak-to-peak amplitude of the sinusoidal wave along the sawtooth. For the hypothetical reference circuit this ratio should not be less than 0·80.

Using the same test signal the amplitude of the synchronizing signal should lie within the limits 0·21 and 0·33 volt when the gain of the circuit is 0 dB, irrespective of whether the three intermediate lines in the test signal are at the black or level.

(2.3.2.5) *Linear Waveform Distortion.*

C.C.I.R. Recommendation No. 267 classifies the perm linear waveform distortion under the headings of 'field- 'line-time' and 'short-time' distortions. Since the test s and the tolerances differ for 405-, 525-, 625- and 819-line sy reference should be made to the original text for details.

In the case of 405-line systems Annex IV of the Recomm tion describes 'acceptance test' and 'routine test' specific based on the use of sine-squared pulse and half-line ba signals (see Reference 16). The sine-squared pulses have amplitude durations of 0·17 microsec (a half period of 3 Mc/s) or 0·33 microsec (a whole period of $f_c = 3$ distinguished by the terms 'T-pulse' and '$2T$-pulse' respec the former being used in acceptance tests and the lat certain of the less critical routine tests. A 50 c/s square test signal associated with line-synchronizing pulses is also The permissible tolerances on the received video waveform defined[16] in terms of a rating factor K.

(2.4) Hypothetical Reference Circuits and Circuit Nois Television Radio-Relay Systems

The signal/noise ratio requirements for television radio systems, as in the case of f.d.m. telephony radio-relay sy are critical in determining the design and cost. In consi the interpretation of the Recommendation relating to signa dom-noise ratios summarized in Section 2.3.2.2, the C. has taken into account the following factors:

(i) For design purposes it is desirable to specify the signa ratio for percentages other than 1% of a month (to whi signal/noise ratio X in Table 1 applies).

(ii) It may be necessary, to allow for fading, to accept signal/noise ratios than X for less than 1% of the time; on th hand signal/noise ratios higher than X are possible for n the time.

(iii) Since the relative distribution with time on radi systems for television and f.d.m. telephony is similar, it is priate to adopt similar methods for specifying the signal/nois performance (Section 2.2).

The C.C.I.R. has recommended (Recommendation No that in the 2 500 km hypothetical reference circuit for the mission of television the ratio of peak-to-peak signal, exc synchronizing pulses, to the r.m.s. value of continuous ra noise read on an instrument having an effective time-co (or integrating time) in terms of power of 1 sec, and u specified weighting network, should not fall below values below in terms of X, where X is the appropriate figure from Table 1:

(*a*) $(X + 4)$ decibels for more than 20% of a month wh fading is severe.

(*b*) $(X - 8)$ decibels for more than 0·1% of a month fading is severe.

The values given are provisional.

It is also recommended that, in part of a hypothetical re circuit consisting of one or two of the three equal homog sections shown in Fig. 6, the mean noise power which sho be exceeded for more than 20% of a month shall be con to be proportional to the number of homogeneous s involved, and that the small percentages of a month which the signal/noise ratio may fall below $(X - 8)$ d shall be regarded as proportional to the number of homog sections involved.

The foregoing recommendations are design objecti

-relay systems with line-of-sight paths. Recommendations ing to other systems, namely those using tropospheric-scatter agation, are referred to in Section 8.6.

2.5) Multi-Channel Voice-Frequency Telegraphy Systems

ne telephone circuits provided by radio-relay systems may be used to transmit multi-channel voice-frequency (v.f.) raphy. The arrangement generally used comprises a block 2, 18 or 24 a.m. or f.m. tones in each audio channel, each tele- h channel operating at 50 bauds. The conditions to be when telephone circuits are to be used to transmit voice- uency telegraphy have been defined by the C.C.I.F. in terms ne permissible attenuation distortion, variations of level with , phase distortion, frequency stability and crosstalk.[12] A ussion of some of the factors affecting radio-relay system gn is given in Report No. 132.

ne level of noise permissible in the case of v.f. telegraphy is g studied by the Joint C.C.I.T.T./C.C.I.R. Working Party Circuit Noise. A provisional recommendation (not yet oved by a Plenary Assembly), which applies to the 2 500 km othetical reference circuit, is that the mean unweighted noise er measured or calculated with a time-constant (integrating) of 5 millisec, and referred to a zero-level point, should not ed 10^6 pW for more than 0·001% (10^{-5}) of any month or more than 0·1% of any hour. It is recognized that sub- sion in terms of circuits shorter than the 2 500 km hypo- ical reference circuit and the use of a time-constant of illisec may present difficulties in practice in view of the very ll percentages of time involved (i.e. less than 0·001% of month). Provided that short bursts of high-level noise due auses other than propagation are negligible and that the cture of the noise is the same as that of white noise, it is sidered by the Working Party that the following objective be used for design purposes:

The unweighted noise power in a telephone channel at a zero lative level point, calculated from measurements made with an tegration time of 1 sec, should not exceed 2×10^5 pW for more an 10^{-4} (i.e. for more than 0·01%) of any month.

With regard to the objective to be met in any hour, it may pen that unforeseen exceptional propagation conditions may ult in this objective not being met during certain most avourable hours. These hours, called 'hours of interrupted graph traffic', will be those during which a noise level of pW is exceeded for more than 36 sec. It follows from the ommended objective for telephony signalling (Section 2.2) t there should never be more than 7 'hours of interrupted graph traffic' in a month. Even so, the total time a noise el of 10^6 pW is exceeded should be not more than 7×36 sec, 4·2 min in a month.

3) PREFERRED BASEBAND CHARACTERISTICS (F.M. SYSTEMS)

(3.1) F.D.M. Multichannel Telephony Systems

t is convenient to define the preferred baseband characteristics .d.m. multichannel telephony systems for systems with various acities, i.e. maximum numbers of telephone traffic channels, ging from 24 to 1 800 as shown in Table 2 (Recommendation . 269). The capacities chosen correspond to integral numbers groups or supergroups and are, in most cases, the same as for e or coaxial-cable systems in current use in order to facilitate erconnection between radio-relay and line systems. The ferred characteristics shown in the Table are appropriate to e-of-sight and near line-of-sight systems, and also to tropo- eric-scatter systems.

It will be noticed that the frequency limits of the baseband in some cases exceed those of the telephone traffic channels; this is because certain line pilots may be associated with the telephony channels for gain regulation, frequency control, etc., and must be transmitted freely by the radio-relay system. In addition, some extension of the baseband beyond the frequency limits shown in the Table may be necessary in order to transmit continuity pilots and provide noise-monitoring channels peculiar to the radio-relay system (Section 6.2).

Table 2

PREFERRED BASEBAND CHARACTERISTICS

F.D.M. MULTICHANNEL TELEPHONY

Maximum number of telephone traffic channels	Limits of band occupied by telephone channels	Frequency limits of baseband	Nominal impedance at baseband	Relative power level per channel*	
				Input	Output
	kc/s	kc/s	ohms	dB	dB
24	12–108†	12–108	150 (bal.)	−45	+ 4·5
60	12–252	12–252	150 (bal.)	−45	+ 1·75
	60–300	60–300	75 (unbal.)	−45	−15
120	12–552	12–552	150 (bal.)	−45	+ 1·75
	60–552	60–552	75 (unbal.)	−45	−15
300	60–1 300 64–1 296	60–1 364	75 (unbal.)	−45	−15
600	60–2 540 64–2 660	60–2 792	75 (unbal.)	−45	−20
960	60–4 028 316–4 188‡	60–4 287	75 (unbal.)	−45	−20
1 800	312–8 204 316–8 204	300–8 248	75 (unbal.)	−40	−25

* Referred to a point of zero relative level.
† For 12-channel systems either the basic group 12–60 kc/s or 60–108 kc/s recommended by the C.C.I.T.T. may be accommodated in the band 12–108 kc/s.
‡ Although not referred to specifically in Recommendation No. 269, it is to be noted that the band 316–4 188 kc/s is suitable only for up to 900 telephone channels, i.e. three mastergroups.

The power levels defined in the last two columns of the Table are the levels at the input and output of the radio-relay system relative to the level of an audio tone applied on the 2-wire side of a 2/4-wire junction which forms part of each telephone circuit.

If, as is often the case, a test tone of 1 mW is applied at such a point, which is referred to as a point of 'zero relative level', the levels shown in the Table correspond to test-tone levels relative to 1 mW. As can be seen from the f.d.m. signal levels given in Section 2.2(*d*), the speech mean power level is appreciably less than the test-tone level.

(3.2) Television Systems

The preferred characteristics for baseband interconnection points of television radio-relay systems are as given in Section 2.3.1.

(3.3) Television and F.D.M. Multichannel Telephony on the Same Radio Carrier

In the case of the largest-capacity radio-relay systems at present defined, e.g. the system for 1 800 telephone channels referred to in Table 2, it is considered that such systems may be used alternatively for the transmission of a television channel and a number of f.d.m. telephony channels simultaneously on each radio carrier. Such an arrangement offers economic advantages compared with systems requiring two radio carriers to provide the same capacity. For the reasons referred to in Section 1.6.2, it is preferred to transmit the video signal in the lower part of the baseband and the f.d.m. multichannel telephony signals in the upper part of the baseband, as shown for example in Fig. 7 (Recommendation No. 271). It is to be noted that,

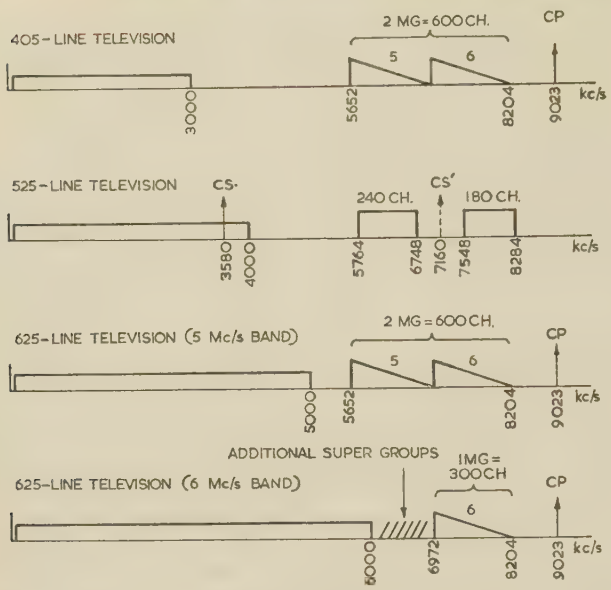


Fig. 7.—Examples of alternative baseband arrangements for radio-relay systems transmitting television and telephony simultaneously on the same r.f. carrier.

The mastergroup numbers shown are those proposed by the C.C.I.T.T.

MG = Mastergroup of 300 telephone channels.
CP = Continuity pilot.
CS = Colour sub-carrier.
CS′ = Second harmonic of CS.

in systems transmitting colour-television signals with a colour sub-carrier, certain blocks of telephone channels may have to be omitted in order to avoid possible interference from the second harmonic of the colour sub-carrier produced by residual non-linearity of the radio-relay system.

(3.4) Television and Sound Channels on the Same Radio Carrier

It is sometimes required to transmit a sound channel (music circuit) simultaneously with a video signal on the same radio carrier. Such an arrangement is often more economical and may have operational advantages compared with the use of a separate land line for the sound channel. A convenient means for providing such simultaneous transmission is to use an f.m. sub-carrier located in the baseband above the video signal but below the radio-relay-system pilot carrier. The frequency and modulation characteristics of the sub-carrier must be carefully chosen so as to avoid crosstalk from the video channel into the sound channel and vice versa, and to achieve a satisfactory signal/noise ratio in the sound channel when the video signal/noise ratio is satisfactory. The preferred characteristics (Recommendation No. 272) are as follows:

(*a*) Bandwidth of sound channel: 30 c/s to 10 kc/s (the upper limit may be increased to 15 kc/s, should this be recommended by the C.C.I.T.T.).
(*b*) Audio level and impedance: +9 dBm in a circuit of 600 ohms balanced.
(*c*) Nominal frequency of sub-carrier: 7·5 Mc/s.
(*d*) Nominal deviation of sub-carrier: 140 kc/s r.m.s. for audio level of +9 dBm.
(*e*) Amplitude of the unmodulated sub-carrier should be such as to produce a deviation of the i.f. and r.f. carriers of 300 kc/s r.m.s.

Pre-emphasis of the frequency-modulated sub-carrier is not considered desirable in view of the possibility of excessive crosstalk into the video channels.

Since operational requirements often demand a number of sound channels, consideration is to be given to preferred characteristics for arrangements whereby up to six sound channels be transmitted, e.g. by means similar to the above.

(4) PREFERRED INTERMEDIATE FREQUENCY AND MODULATION CHARACTERISTICS (F.M. SYSTEMS)

Where microwave radio-relay systems are to be establi across international frontiers it is obvious that the modula characteristics of the r.f. carrier, including the frequency de tion of the carrier and the use or otherwise of pre-emph must be agreed by the countries concerned. From thi follows that the modulation characteristics of the i.f. carrier also determined. The need for a preferred intermediate quency is perhaps less obvious, and it arises from the follo considerations:

(*a*) At common frontier stations international interconnectio be made at intermediate frequency without translation to baseb thus reducing noise and distortion.
(*b*) Within a country the use of a standard intermediate frequ facilitates interconnection between national systems.
(*c*) Television outside-broadcast injection facilities can be r readily provided.
(*d*) Maintenance and testing of systems is simplified.
(*e*) Preparation of preferred r.f. channel arrangements (Secti is facilitated.

Except where otherwise shown the preferred i.f. characteri discussed below apply to line-of-sight and near line-of-s radio-relay systems with capacities of up to 1 800 telep channels or the equivalent; characteristics suitable for o systems such as tropospheric-scatter systems have yet t determined, although those shown below may be used w appropriate.

(4.1) Intermediate Frequency Characteristics (Recommenda No. 273)

(4.1.1) Centre Value of the Intermediate Frequency.

The 'centre value' of the intermediate frequency is that co sponding to the arithmetic mean of the upper and lower li of the deviation of the carrier; in most systems this corresp closely to the frequency of the carrier in the absenc modulation, but in some cases, e.g. television systems wit pre-emphasis, the intermediate frequency in the absenc modulation may be offset relative to the centre value.

The preferred nominal centre values of the intermediate quency, which apply both to multichannel telephony (f.d systems and to television systems, are as follows:

(i) 35 Mc/s for radio frequencies up to 1 000 Mc/s (70 Mc/s also be used, for example, in larger-capacity radio-relay system
(ii) 70 Mc/s for radio frequencies from 1 000 to 8 500 Mc/s.

(4.1.2) Input and Output Voltages and Impedance of the Interme Frequency Circuit.

In order to facilitate interconnection at intermediate frequ the following preferred values have been defined:

(*a*) Output voltage: 0·5 volt r.m.s.
(*b*) Input voltage: 0·3 volt r.m.s.

The difference between these voltages allows for losses i connecting cables.

(*c*) Nominal impedance: 75 ohms unbalanced.

In the case of f.d.m. systems the amount of intermodul noise due to mismatch of the i.f. circuits at a point of i national interconnection is required not to be excessiv relation to the overall noise requirements of the hypoth reference circuit for such systems.

Similarly, in the case of television systems the echo disto due to impedance mismatch of the i.f. circuits at a point of national interconnection is required not to be excessiv

ion to the overall performance requirements of the hypo-
cal reference circuit for such systems.

.2) Modulation Characteristics for F.D.M. Multichannel Telephony

ne modulation characteristics given below apply to the i.f.
r.f. carriers of f.m. systems for f.d.m. telephony.

l) Frequency Deviation without Pre-Emphasis.

ne choice of an optimum value for the frequency deviation
.d.m. multichannel telephony is conditioned by a number of
nical and economic factors. Larger values of the deviation
le the signal/basic-noise ratio to be improved, or if the
l/noise ratio is unchanged, the cost of a system to be
ced by using smaller aerials, lower-power transmitters or
er paths. On the other hand, large deviations may increase
ly the level of intermodulation noise arising from i.f. and
delay distortion, feeder echoes or multi-path transmis-
[13, 14] Large deviations are also undesirable in view of the
ased bandwidth that would be needed and the resulting
of economy in the use of the spectrum. Consideration of
e factors has resulted in the preferred values shown in
e 3 for the carrier-frequency deviation per telephony channel
systems not using pre-emphasis (Recommendation No. 274).

Table 3

PREFERRED CARRIER-FREQUENCY DEVIATION (WITHOUT PRE-EMPHASIS)

F.D.M. MULTI-CHANNEL TELEPHONY

Maximum number of telephone traffic channels	R.M.S. deviation for a single telephone channel*
	kc/s
24	35
60	50, 100, 200 alternative
120	50, 100, 200 values
300	200
600	200
960	200
1 800	140†

* For a tone of 1 mW at a point of zero reference level.
† Provisional value, with pre-emphasis.

2) Pre-Emphasis Characteristics

is well known that, in the absence of pre-emphasis, the
c noise in the baseband output of an f.m. system has a
ngular' distribution, i.e. the noise voltage in a given narrow
d is proportional to the centre frequency of that band.
wever, noise from modulators and demodulators may modify
triangular noise distribution in practice. Thus in f.d.m.
phony systems the signal/basic-noise ratio in the telephony
nnels varies in accordance with the position of the channel
he baseband and some channels may have appreciably better
ial/noise ratios than are necessary. It is reasonable, there-
, to attempt to obtain a uniform signal/noise ratio in all the
nnels, and a more economical system, by increasing the
rier-frequency deviation for the upper channels relative to
lower channels. However, if this process is carried too far
level of intermodulation noise in the lower-frequency channels
omes excessive. Such intermodulation noise arises from non-
arity of modulators and demodulators, from delay distortion
ntermediate and radio frequencies, as noted earlier, and as a
ult of feeder echoes and multi-path effects. A compromise is
refore necessary, and the preferred pre-emphasis charac-
stic shown in Fig. 8 has been adopted for f.d.m. multichannel

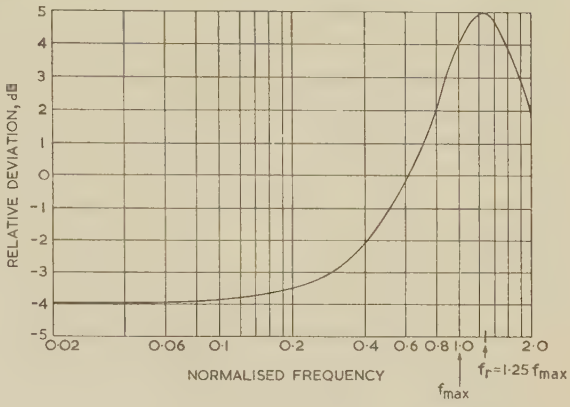


Fig. 8.—Pre-emphasis characteristic for f.d.m. telephony.

telephony systems with capacities of up to 960 channels, and provisionally for systems of 1 800 channels.

The preferred pre-emphasis characteristic (Recommendation No. 275) is that obtained by using a network with an insertion loss given by

$$20 \log_{10} \frac{E}{V_1} = 10 \log_{10} \left[1 + \frac{6 \cdot 90}{1 + \frac{5 \cdot 25}{\left(\frac{f_r}{f} - \frac{f}{f_r} \right)^2}} \right] \text{decibels}$$

where E and V_1 are the input e.m.f. and output voltage, respectively, f_r is the resonant frequency of the network ($f_r = 1 \cdot 25 f_{max}$), and f_{max} is the highest telephone-channel baseband frequency of the system.

The basic pre-emphasis and de-emphasis networks are shown in Fig. 9, and values of f_{max} and f_r for f.d.m. systems of various capacities are given in Table 4.

Table 4

CHARACTERISTIC FREQUENCIES FOR PRE- AND DE-EMPHASIS NETWORKS

F.D.M. MULTICHANNEL TELEPHONY

Maximum number of telephone traffic channels	f_{max}	f_r
	kc/s	kc/s
24	108	135
60	300	375
120	552	690
300	1 300	1 625
600	2 660	3 325
960	4 188	5 237†
1 800	8 204*	10 255*

* Provisional values; the networks are also provisional for 1 800-channel systems.
† This value, given in Recommendation No. 275, appears to be incorrect and should be 5 235 kc/s.

It is an additional requirement that where pre-emphasis is used the effective (r.m.s.) deviation due to the multichannel signal should be the same with and without pre-emphasis. This requirement ensures that the bandwidth occupied by the modulated r.f. and i.f. carriers shall be the same with and without pre-emphasis, thus enabling the same r.f. channel arrangement to be used in both cases.

When setting up the desired modulation conditions it is useful to note that the deviation for a single telephone channel, with pre-emphasis, is the same as that without pre-emphasis for the baseband telephone channel whose centre frequency is $0 \cdot 608 f_{max}$.

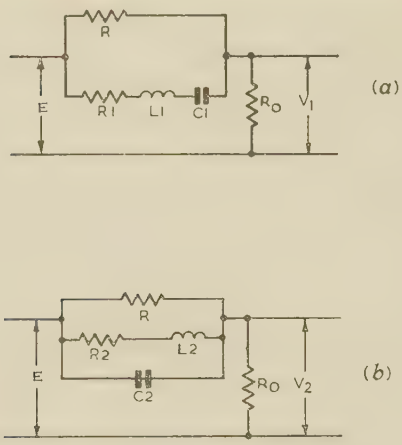


Fig. 9.—Basic pre-emphasis and de-emphasis networks for f.d.m. multi-channel telephony.

(*a*) Basic pre-emphasis network

$$R = 1{\cdot}81R_0$$
$$R_1 < 0{\cdot}01R_0 \text{ at } f_r$$
$$\sqrt{\frac{L_1}{C_1}} = 0{\cdot}79R_0$$
$$f_r = 1{\cdot}25f_{max} = \frac{1}{2\pi}\sqrt{\frac{1}{L_1C_1}},$$

where f_{max} is the highest baseband frequency.

(*b*) Basic de-emphasis network

$$R = 1{\cdot}81R_0$$
$$R_2 < 0{\cdot}02R_0 \text{ at } f_r$$
$$\sqrt{\frac{L_2}{C_2}} = 1{\cdot}47R_0$$
$$f_r = 1{\cdot}25f_{max} = \frac{1}{2\pi}\sqrt{\frac{1}{L_2C_2}}$$

Element tolerances for networks (*a*) and (*b*) are as follows:

Resistors : ±1%
Capacitors: ±0·5%
f_r : ±0·5% (*L* resonating with *C*)

(4.3) Modulation Characteristics for Television

The modulation characteristics given below apply to the i.f. and r.f. carriers of f.m. systems for television.

(4.3.1) Frequency and Sense of Modulation.

The technical and economic factors that determine the optimum frequency deviation for television are similar to those for f.d.m. multichannel telephony discussed earlier; increasing the deviation improves the signal/noise ratio, but if taken too far it may not be justified because of the increased spectrum space required. Furthermore, large deviations may produce excessive waveform distortion and non-linearity, particularly at high baseband frequencies. Such distortion and non-linearity are troublesome in the transmission of colour-television signals since they give rise to differential amplitude and phase distortion.

The preferred values of the deviation for the peak-to-peak amplitude of the video signal (Recommendation No. 276) are as follows:

625 lines or less	8 Mc/s peak-to-peak
819 lines	8 to 12 Mc/s* peak-to-peak

* The actual value is to be agreed between the administrations concerned.

The above values apply both without and with the pre-emphasis characteristics discussed below.

The 'sense' of the modulation, i.e. whether a positive-going video signal corresponds to an increase or a decrease of the radio or intermediate frequency, has not been standardized, since it is a relatively easy matter to reverse the polarity of the video signal as required. Furthermore, in some systems the sense of the modulation is intentionally reversed on succe sections of a multi-section link in order to minimize distor

(4.3.2) Pre-Emphasis Characteristics.

The use of pre-emphasis in television transmission is of siderable value for the following reasons:

(*a*) It transforms a non-symmetrical waveform to one of g symmetry (see Fig. 10), thereby enabling the same radio-relay sy from the modulator to the input demodulator output, to be alternatively for television and f.d.m. multichannel telephony.

(*b*) Frequency stabilization of the r.f. and i.f. carriers is simp since the same mean-frequency control can be used for tele and f.d.m. multi-channel telephony.

(*c*) Distortion, including differential amplitude and phase d tion, is minimized, thus facilitating the transmission of c television signals or of sound by means of a sub-carrier.

The optimum shape of the pre-emphasis characteris determined by several factors. A large reduction in amp of the low-frequency components of the video signal is sirable since it may result in an unsatisfactory signal/interfe ratio due to mains hum, microphony and other low-freq

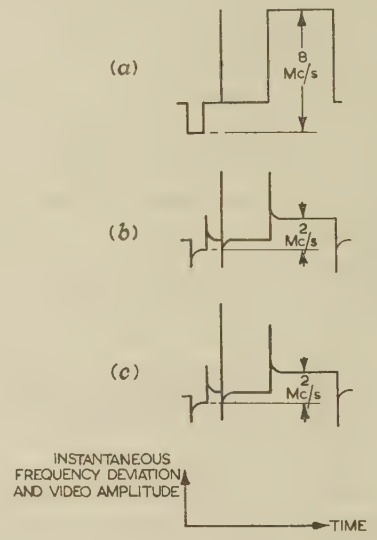


Fig. 10.—Effect of pre-emphasis on video waveform and instantaneous carrier-frequency deviation.

(*a*) *T* or 2*T* pulse and bar without pre-emphasis.
(*b*) 2*T* pulse and bar with pre-emphasis.
(*c*) *T* pulse and bar with pre-emphasis.

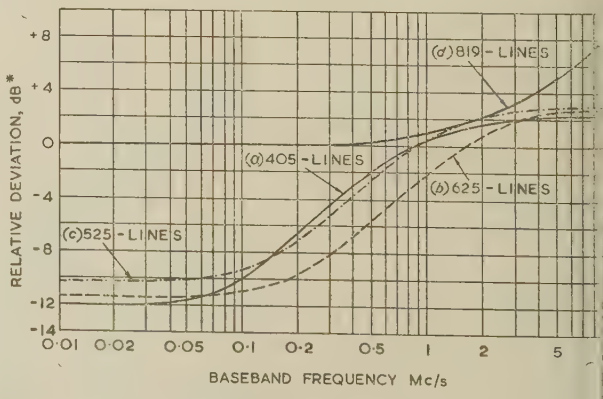


Fig. 11.—Pre-emphasis characteristics for 405-, 525-, 625- a 819-line television systems.

* 0 dB corresponds to a deviation of 8 Mc/s for 1-volt peak-to-peak si signal.

Table 5

PREFERRED R.F. BANDS FOR INTERNATIONAL RADIO-RELAY SYSTEMS

Maximum capacity of each radio carrier (telephone channels or television)	Nominal limits of radio frequency band	Preferred 'centre' frequency f_0	Width of r.f. band occupied
	Mc/s	Mc/s	Mc/s
60–120 (Recommendation No. 283) ..	1 700–1 900 } (2 000 Mc/s band)	1 808	200
	1 900–2 100 }	2 000	200
	2 100–2 300 }	2 203	200
60–120 (Recommendation No. 284) ..	7 425–7 725 (7 000 Mc/s band)	7 557·5	300
300–1 800 or television or equivalent (Recommendations Nos. 278, 279 and 281)	1 700–2 100 } (2 000 Mc/s band)	1 903	400
	1 900–2 300 }	2 101	400
	3 800–4 200* (4 000 Mc/s band)	4 003·5	400
600–1 800 or television or equivalent (Recommendations Nos. 280 and 281)	5 925–6 425 (6 000 Mc/s band)	6 175	500

* In Regions 2 and 3 the band 3 700–4 200 Mc/s is used.

rference. A large increase in the amplitude of the high-uency components of the video signal is also undesirable e it produces large 'spikes' when rapid video transients ır (see Fig. 10), and these may cause interference in adjacent nnels. Furthermore it is desirable for the signal/thermal-ıe ratio to be substantially unchanged as a result of the oduction of pre-emphasis. These considerations have led to adoption of the preferred pre-emphasis characteristics for -, 525-, 625- and 819-line television systems shown in Fig. 11. se characteristics can be achieved by the minimum phase-t pre-emphasis network in Fig. 12 (Recommendation

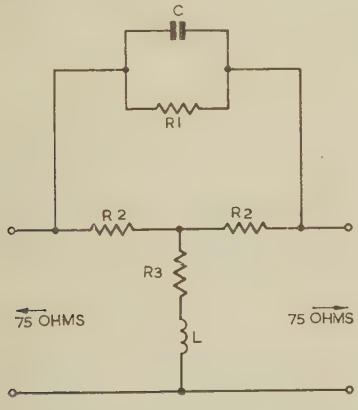


Fig. 12.—Basic pre-emphasis networks for 405-, 525-, 625- and 819-line television systems.

Curve in Fig. 11	(*a*)	(*b*)	(*c*)	(*d*)
Lines	405	625	525	819
L, μH	22·25	9·54	15	3·313
C, pF	3950	1695	3085	589
R_1, ohms ..	300	300	∞	121·3
R_2, ohms ..	75	75	48·7	75
R_3, ohms ..	18·75	18·75	33·2	46·38

·. 277), the corresponding de-emphasis network being the erse of the pre-emphasis network.

PREFERRED FREQUENCY BANDS AND R.F. CHANNEL ARRANGEMENTS (F.M. SYSTEMS)

The following discussion relates to the frequency bands and channel arrangements for radio-relay systems using line-of-ht or near line-of-sight paths.

In order that radio-relay links may be established across national frontiers it is clearly essential that there should be agreement between the administrations directly concerned on the frequency bands and r.f. channel arrangements to be used. Such agreement has been greatly facilitated by the international adoption of certain preferred characteristics.

The broad frequency bands allocated to the fixed radio service have been laid down in the Radio Regulations adopted by the Administrative Radio Conference (Geneva, 1959). The C.C.I.R. has recommended the use of certain preferred bands within these broad allocations for international radio-relay links, and these bands have also been widely used for national radio-relay links. The preferred frequency bands for multi-channel telephony systems of various capacities and for television systems are shown in Table 5.

The r.f. channel arrangements consist of symmetrical patterns related to the nominal 'centre' frequencies shown in Table 5. However, for reasons given below, the 'centre' frequencies are not precisely located at the middles of the corresponding frequency bands.

The choice of preferred r.f. channel arrangements for systems of various capacities is determined by several factors and is a matter of compromise. The more important factors concerned are as follows:

(i) Several r.f. channels may be required along the same route and provision may be needed for r.f. channels on crossing or spur routes.

(ii) For reasons of economy and convenience it may be desired to use common transmit-receive aerials (as shown, for example, in Fig. 1).

(iii) The spacings between the channels and their arrangement must be such as to avoid mutual interference, without requiring filters of excessive complexity and cost.

(iv) Due regard must be paid to economy in the use of the spectrum.

The basic pattern generally adopted is one in which the transmit (or receive) frequencies are in one block and the receive (or transmit) frequencies are in a second block alongside the first. This arrangement simplifies filtering by increasing the spacing of the high-level transmit and low-level receive signals. Figs. 13(*a*)–(*c*) and (*e*) show r.f. channel arrangements for the radio-relay systems given in Table 5. However, in some systems a pattern of alternate transmit and receive channels has been adopted [as in Fig. 13(*d*), which shows the r.f. channel arrangement used in the American TD2 system].[4] For reasons of economy in use of the spectrum a '2-frequency' per channel pattern is usually employed, i.e. a given channel uses f_n on section 1 of the route, f_n' on section 2, f_n again on section 3 and

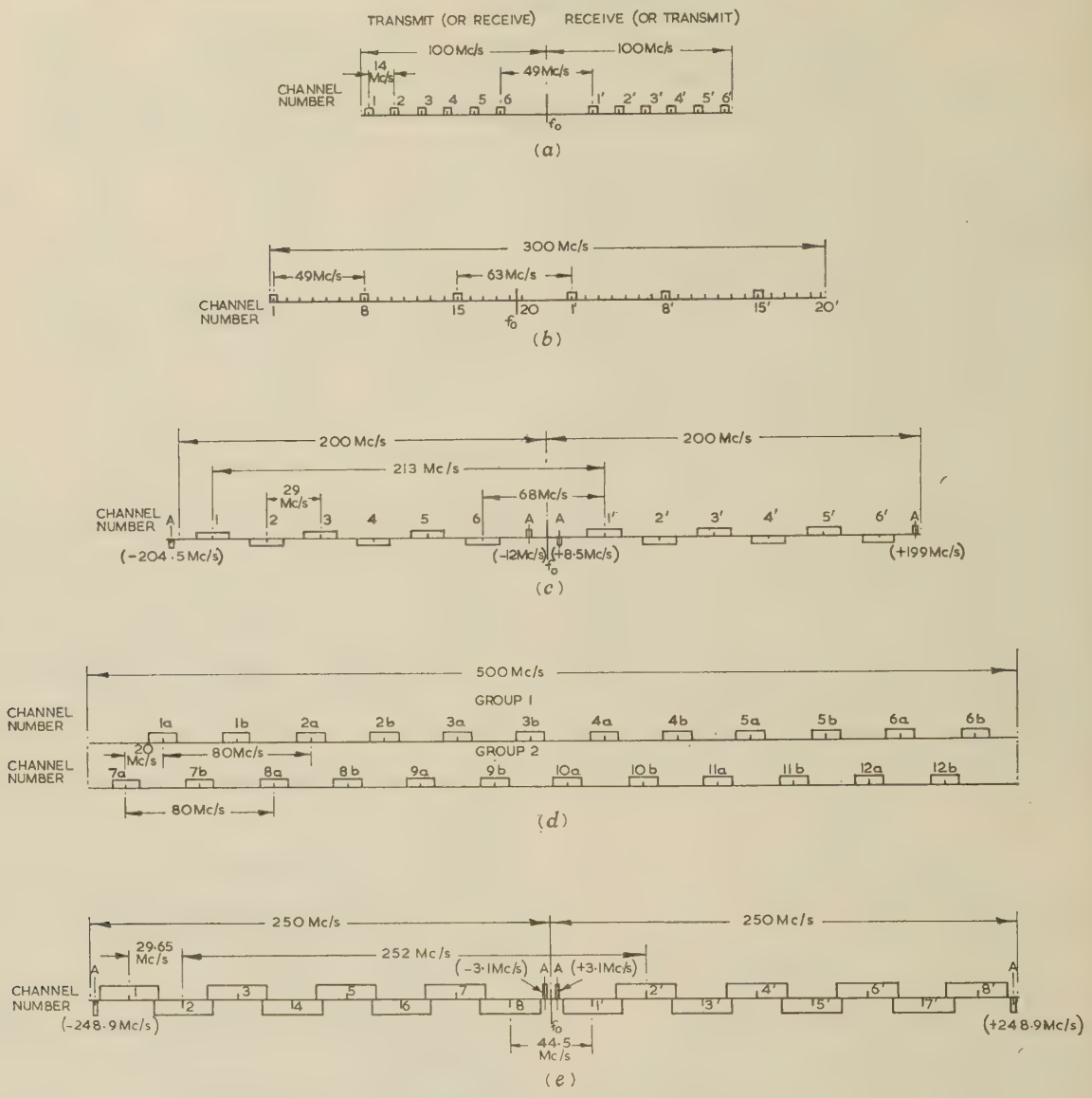


Fig. 13.—Preferred r.f. channel arrangements.

(*a*) 60- or 120-channel f.d.m. systems operating in the 2 000 Mc/s band.
(*b*) 60- or 120-channel f.d.m. systems operating in the 7 000 Mc/s band.
(*c*) 600- to 1 800-channel f.d.m. systems or the equivalent operating in the 2 000 and 4 000 Mc/s band.
A = Auxiliary radio-relay system channels.
▭ or ▭· indicates H(V) or V(H) polarization)

(*d*) Alternative to (*c*) for 600-channel f.d.m. systems or the equivalent operating in the 4 000 Mc/s band (Regions 2 and 3).
(*e*) 600- to 1 800-channel f.d.m. systems or the equivalent operating in the 6 000 Mc/s band.
A = Auxiliary radio-relay system channels.

so on, the difference $f'_n - f_n$ being the 'shift' frequency. It is usual to employ different polarizations on alternate sections, e.g. sections 1 and 3, using the same frequency. However, in some situations where 'overreach' interference is liable to occur, e.g. between sections 1 and 3 due to insufficient aerial discrimination, it may be necessary to use a slightly different frequency on section 3. Different frequencies may also be needed on corresponding channels for different directions of transmission on adjacent sections, in order to avoid interference due to foreground reflections. Provision is therefore made for the use of 'stagger' or interleaved frequencies where necessary, these being midway between the frequencies of the basic pattern. (In the 60–120-channel systems operating in the 7 000 Mc/s band quarter-spacing stagger frequencies are also available).

There are several possible sources of internal interferenc multi-r.f.-channel systems, and great care is necessary if preparing r.f. channel arrangements, such interference is t avoided without giving rise to filtering difficulties. Interfe may arise from the following causes:

(*a*) Image frequencies of the transmitter or receiver (in sys with an intermediate frequency).

(*b*) Intermodulation products of two or more transmit sig due, for example, to amplitude non-linearity at joints in wavegu

(*c*) Signals which fall near the local-oscillator frequency o receiver or twice the intermediate frequency from it.

(*d*) Harmonics of the shift frequency $f'_n - f_n$.

The choice of an r.f. channel arrangement which minim such interference has been described elsewhere, with partic reference to the arrangement shown in Fig. 13(*c*) for 600–1

nel f.d.m. systems, or the equivalent, operating in the 2000 4000 Mc/s bands.[17]

ne centre frequencies shown in Table 5 and indicated by f_0 ig. 13 have been selected to minimize certain of the sources terference discussed above—in particular, those related to nonics of the shift frequency $f'_n - f_n$.

l the r.f. channel arrangements shown in Fig. 13 are suitable se with the preferred intermediate frequency of 70 Mc/s. arrangement shown in Fig. 13(*e*) is also suitable for an mediate frequency of 74·13 Mc/s, which enables a common lator (14·82 Mc/s) to be used for generating all the local lations of the system (this scheme is used in the American system).[5]

here common transmit-receive aerials are employed it is ded that only suitably spaced channels, e.g. the alternate nels in Figs. 13(*c*) and (*e*), shall be employed on the same l and that these channels shall use the same polarization. remaining channels could be used, if desired, on a separate smit-receive aerial with a different polarization.

(6) SUPERVISORY, CONTROL AND MAINTENANCE TECHNIQUES

eliability is an essential requirement for both international national circuits. The standards of reliability aimed at are s that the outage time due to all causes, including adverse pagation, equipment and mains supply failures, should not ed some 0·05% per annum for a 2500 km circuit, and portionally less for shorter circuits. It is also necessary to ntain the desired standard of transmission performance, i.e. signal/noise ratio should not fall below, or the distortion ed, the prescribed limits.

he achievement of high reliability and the maintenance of sfactory transmission performance require the following lities:

(*a*) Supervisory facilities to enable the overall continuity and erformance of the broadband channels, the correct functioning of e equipment and availability of the mains supply at both the ttended and unattended stations to be continuously and auto- natically monitored and alarms given at attended control stations n the event of failure.

(*b*) Control facilities to enable the appropriate switching, e.g. rom working to protection channels or from mains to standby ower supplies, to be carried out either automatically or manually.

(*c*) Service channels for the transmission of maintenance informa- ion between staff at attended control stations, and to staff visiting ormally unattended stations for fault clearance; service channels re also needed for the transmission of control and supervisory ignals between attended and unattended control stations.

(*d*) Testing facilities for the examination of the overall (baseband o baseband) transmission performance of the channels (the facilities equired for detailed testing at individual stations are excluded from his discussion since they do not normally involve international o-operation).

The techniques used to achieve high reliability, maintain the nsmission performance and locate faults are described below. the case of international circuits it is evident that common hniques in the countries concerned are desirable, if not ential, and this has led to a certain degree of international ndardization.

The overall aspects of the maintenance of international radio- ay and line systems are the responsibility of the C.C.I.T.T. ce they concern the performance of the general network, but detailed aspects of the maintenance of international radio- ay systems are the responsibility of the C.C.I.R.

(6.1) Control Stations, Switching Sections and Protection Channels

It is the practice to subdivide long radio-relay systems into itched sections 150–300 km in length, each switched section comprising up to six or so unattended intermediate (repeater) stations. The switched sections terminate in control stations which are either attended or operated remotely from nearby attended stations. Over the switched sections a number of working channels are provided to meet the traffic requirements, together with a protection channel which can be switched automatically or manually into circuit if a working channel fails, as described in Section 1.5. The criterion adopted for switching is an excessive change in level of the continuity pilot or an abnormal increase of noise in a noise-measuring channel (Section 6.2). Where the characteristics of the working channels are sufficiently alike, e.g. for television and 960-circuit multi-channel telephony, a common protection channel is often used (Recommendation No. 305). A typical arrangement of the working and protection channels, and the associated pilot injection and detection facilities, is shown in Fig. 14.

The use of a switched protection channel is generally preferred to the use of switched standby equipment operating on the same frequency as the working equipment and switched into circuit only at a station where a fault has occurred. The reasons for this preference are as follows:

(*a*) Continuous automatic monitoring of the protection channel can be arranged, thus helping to ensure that the protection channel is ready for service when required.

(*b*) Maintenance is facilitated since a working channel can be replaced by the protection channel and tested end to end without interrupting traffic.

(*c*) The protection channel is available for carrying traffic on exceptional occasions if required, e.g. for television outside-broadcast purposes.

In considering the interruptions due to switching, e.g. from working to protection channels, a distinction is made between the transfer time of the switches and the time taken for the switches to operate after a fault has occurred. The latter is often limited by the time taken to transmit control signals over the switched section. The following objectives for the maximum durations of interruptions have been defined (Report No. 137):

Transfer time.—1 millisec, which appears acceptable for most services. However, if high-speed data transmissions are required, and even for certain types of v.f. (carrier) telegraphy, this value should be considerably reduced.

Operate time.—35 millisec, which may cause errors in 50-baud telegraph transmission.

It is noted, however, that, with systems now in service, the transfer time may be about 3–5 millisec and the operate time may be 50 millisec or more.

It is further recommended that the design of switching systems should be such that interruptions corresponding to the operate time (of the order of 35 millisec) occur only in the case of infrequent and unavoidable failures of the working equipment. Interruptions corresponding to the transfer time (1 millisec or less) may occur more frequently without serious difficulty, e.g. owing to failures in radio propagation arising from selective fading.

(6.2) Pilots and Noise-Monitoring Channels

In order to monitor the continuity or otherwise of the working and protection channels, even when traffic signals are not present, it is convenient to transmit a 'continuity' pilot, usually at a frequency some 10% above the upper frequency limit of the f.d.m. multi-channel telephony signal to facilitate filtering. The precise frequency is chosen, where possible, to minimize intelligible crosstalk in the telephony channels. The continuity pilots for television systems have been selected to be the same as those for f.d.m. multi-channel telephony systems of comparable baseband width. The continuity pilot is individual to

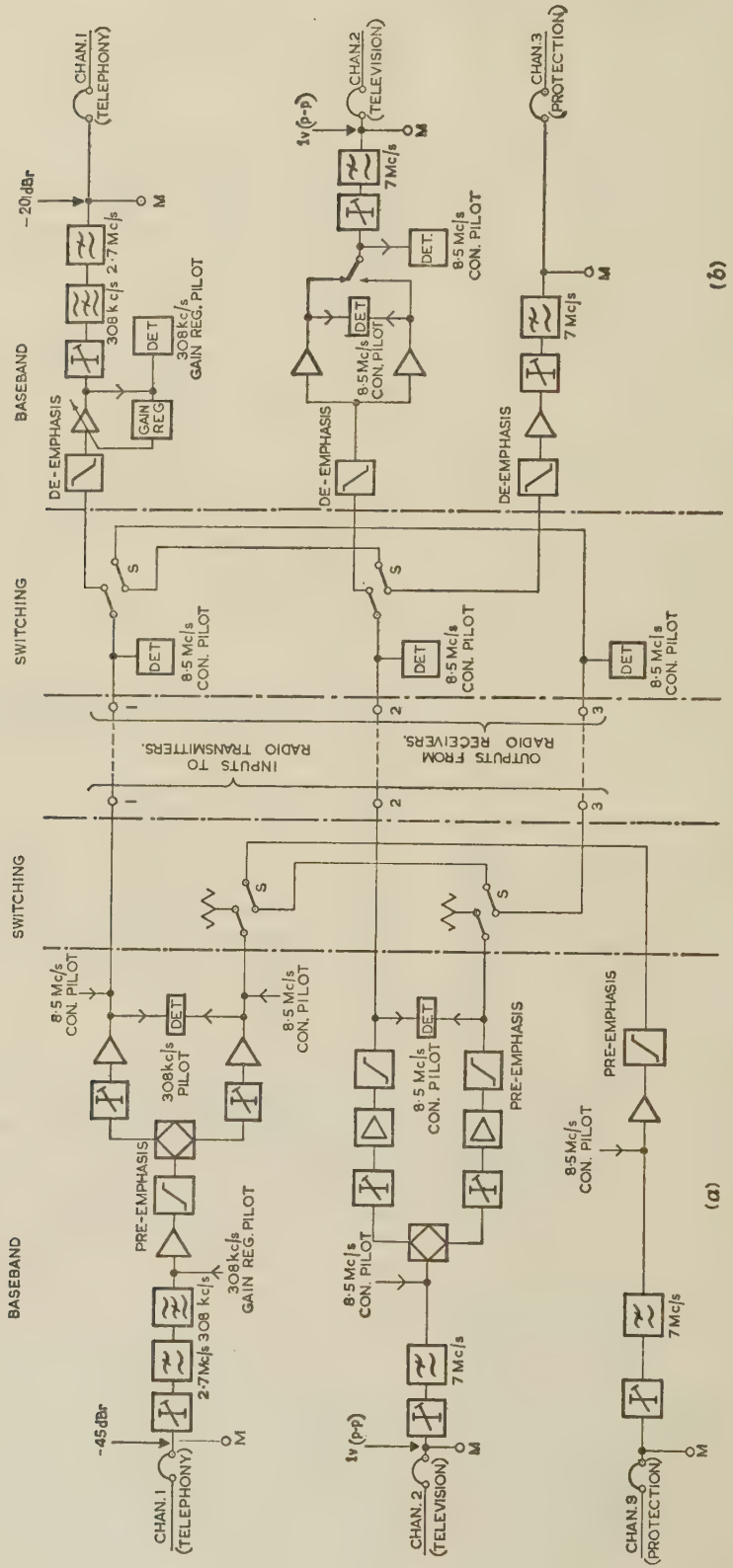


Fig. 14.—Typical arrangement of working and protection channels, continuity and gain-regulating pilots.

M = Monitoring points.
S = Switch operated from continuity pilot detector.
(*a*) Transmit terminal (*b*) Receive terminal.

adio-relay system and is suppressed at the baseband output eans of a band-stop or low-pass filter [Fig. 14(*b*)] to avoid ference to other systems to which the radio-relay system be connected. The preferred frequencies for the continuity s for f.d.m. multi-channel telephony systems of various cities, and for television systems, are shown in Table 6 ommendation No.292).

Table 6

CONTINUITY PILOTS AND NOISE-MEASURING CHANNELS F.D.M. MULTI-CHANNEL TELEPHONY AND TELEVISION

mum number telephone ffic channels television	Continuity pilot frequency	Deviation (r.m.s.) produced by the pilot	Centre frequencies of noise-measuring channels	
			Below traffic channels	Above traffic channels
	kc/s	kc/s	kc/s	kc/s
24	116 or 119	20	10	116 or 119
60	304 or 331	25, 50, 100*	10 or 50	304 or 331
120	607	25, 50, 100*	10 or 50	607
300	1 499, 7 000 or 8 500	100 or 140†	50	1 499
600	3 200 or 8 500	140	50	3 200
960	4 715 or 8 500	140	50 or 270	4 715
1 800	9 023	100	270	9 023
405-line elevision	8 500	140	—	8 500
625-line elevision	8 500	140	—	8 500
levision and up to 600 telephone channels	9 023	100	—	9 023

Alternative values depending on whether the deviation of the signal is 50, 100 or c/s (Table 3).
The first figure applies to a frequency of 1 499 kc/s, and the second to 7 000 or kc/s.

he level of the continuity pilot is most conveniently defined erms of the radio-carrier frequency deviation produced by it. level should not be so low as to create uncertainty due to se present with the carrier, nor so high as to contribute ificantly to the risk of crosstalk or interference in adjacent io channels. The preferred deviations are also shown in ble 6.

he use of a continuity pilot by itself is insufficient, since, ough the pilot may be present, the channel may be unusable ing to excessive noise. It is convenient, therefore, to monitor omatically the noise in a narrow channel, which may be tred on the pilot frequency; alternatively, or in addition, a se-measuring channel below the band occupied by the f.d.m. lti-channel telephony signal may be used. The use of a se-measuring channel centred on the pilot frequency has the vantage of enabling common designs of filters to be used for pilot and the noise-measuring channel, for selection and pression purposes. The preferred centre frequencies of the ise measuring channels are also shown in Table 6 (Recom-ndation No. 293).

In view of the high degree of overall gain stability required by .m. multi-channel telephony systems (Section 2.1.4), it is often sirable to use automatic regulation of the overall gain between seband input and output. It is desirable for this purpose to a gain-regulating pilot within the baseband spectrum of the .m. multi-channel signal, since the level variations of the ot then closely correspond to the unwanted level variations the f.d.m. signal. In the case of f.d.m. systems with a pacity of 60 channels or more the preferred gain-regulating lot frequency is 308 kc/s ± 3 c/s; this pilot frequency is also used for line systems and falls in a convenient gap in the spectrum of the f.d.m. signals (Recommendation No. 291) (see Section 1.6 and Fig. 2). To avoid interference with the gain-regulating pilots of other systems, the radio-relay-system 308 kc/s pilot is required to be suppressed to at least −50 dBm0 at the output of the system.

(6.3) Service Channels

(6.3.1) Types and Functions of Service Channels.

The following recommendations (Recommendation No. 295) have been made with respect to international radio-relay systems capable of transmitting at least 60 telephone channels or a television signal, and comprising two staffed terminal stations, in which the signals are demodulated to baseband, and any number of unstaffed intermediate stations:

(*a*) All staffed stations should be connected directly to the public telephone network.

(*b*) When a radio-relay link is extended by means of short cable sections, and when these cable sections and the radio-relay link taken together constitute a 'regulated line section', the terminal stations of the radio-relay link itself should have speaker circuits to the stations at the ends of the regulated line section.

(*c*) One telephone service channel (a main or express speaker circuit) should be provided for telephonic communication directly between the staffed control stations receiving supervisory signals.

(*d*) A second telephone service channel (an omnibus speaker circuit) should be provided for telephone communication linking together all stations, both staffed and unstaffed, of the system.

(*e*) Wherever possible, and after agreement between the administrations concerned, one or two service channels should be provided in each direction for the transmission of supervisory and control signals between the stations of the system.

(*f*) One of the service channels might be used for the transmission of high-speed signals associated with the switching of broadband radio channels, and the other could be used for the transmission of a number of relatively low-speed supervisory signals.

(*g*) Wherever possible all service channels should provide the bandwidth recommended by the C.C.I.T.T. for international telephone circuits, namely 300–3 400 c/s.

(*h*) The service channels should be provided either over metallic circuits, or over an auxiliary radio-relay system, using either the same frequency band as the main system or a different band, which follows the same route as the main system.

It is also possible, by agreement between the administrations concerned, to transmit the relatively slow signals in the upper part of the omnibus service channel mentioned in (*d*).

In the case of low-capacity systems (maximum of 300 telephone channels) the service channels can, for economic reasons, be placed in the baseband of the main radio-relay systems below the band occupied by the f.d.m. multi-channel telephony signals, by agreement between the administrations concerned.

In connection with (*f*), it is to be noted that, when changing over from a working to a protection channel under fault conditions, the control signals needed to switch both ends of the channel must be transmitted from the receiving end to the transmitting end with a minimum of delay if switching interruptions are to be avoided. The change-over must be initiated from the receiving end since that is the point at which failure of transmission becomes apparent. Other control or supervisory signals, such as those indicating that a mains failure has occurred and that the equipment is operating on standby power supplies, need not be transmitted quickly.

The characteristics of the control and supervisory signals have not so far been standardized, and it is left to administrations to agree on the characteristics to be used.

(6.3.2) Radio-Frequency Channel Arrangements for Service Channels

It is often convenient, for reasons of economy and to avoid adding to the mast loading by additional aerials, to provide the service channels on an auxiliary radio-relay system sharing the same aerials and feeders as the main radio-relay system. The

r.f. channel arrangements for these auxiliary channels must be carefully chosen in order to avoid interference to and from the main r.f. channels, owing to the causes described in Section 5. It is usual to provide two transmit and two receive r.f. channels in order to permit the use of working and protection channels on the auxiliary radio-relay system, with diversity working if required. (The use of frequency diversity in this case is considered to be justified in view of the relatively small additional spectrum space needed.)

The r.f. channel arrangements for auxiliary radio-relay systems operating in the 2000, 4000 and 6000 Mc/s bands are shown in Figs. 13(*c*) and (*e*); corresponding transmit and receive channels have the same polarization whilst the two transmit (or receive) channels have different polarizations (Recommendation No. 296).

(6.4) Testing Techniques for F.D.M. Multi-Channel-Telephony Radio-Relay Systems

The maintenance of a satisfactory standard of performance on f.d.m. multi-channel-telephony radio-relay systems requires the measurement of the following characteristics (Recommendation No. 290):

(*a*) Stability of the overall gain or loss in the baseband.
(*b*) Total noise, including crosstalk or intermodulation noise.

The measurement of the overall (baseband input to baseband output) gain or loss is conveniently carried out by use of the 308 kc/s gain-regulating pilot (Section 6.2) since this can be done without withdrawing the channel from service. Additional measurements of the variation of the gain or loss over the frequency range of the baseband may also be required from time to time, and this normally requires the channel to be withdrawn from service.

A general indication of the transmission quality from the noise aspect can be obtained by observation of the noise in the noise-measuring channels above or below the traffic channels (Section 6.2 and Table 6), and this can be done whilst the channel is in service. More detailed measurements are often required, and for this a white-noise testing signal is preferred, the white-noise signal being used to simulate the loading imposed on the radio-relay system by the active traffic channels. Inserted in the white-noise signal are three or more narrow measurement 'slots' in which the noise level before transmission over the system under test is suppressed by at least 80 dB. These slots correspond approximately to non-active traffic channels. The noise in the measurement channels at the output of the radio-relay system is the sum of the basic (thermal) noise of the system and intermodulation noise arising from non-linearity. This technique is especially valuable for the assessment of intermodulation noise in radio-relay systems, because it enables this to be done simply and directly, whereas measurements of the harmonics or intermodulation products of one or two sinusoidal signals are difficult to interpret because of their variation with time and frequency.[18]

The following characteristics of the white-noise signals have been agreed internationally (Recommendation No. 294):

(i) The power level of the signal with a uniform continuous spectrum should, for radio channels with at least 240 telephone channels, be equal to $(-15 + 10 \log_{10} N)$ decibels relative to 1 mW, at a point of zero relative level, N being the number of traffic channels (see Fig. 15).

(ii) The power level of this signal with a uniform continuous spectrum for radio channels with a capacity of at least 12 and not more than 240 telephone channels should be equal to $(-1 + 4 \log_{10} N)$ decibels* relative to 1 mW at a point of zero relative level, N being the number of traffic channels;

(iii) The frequencies of the measurement channels should be as shown in Table 7.

* Provisional value for systems of less than 60 channels.

Table 7

CHARACTERISTICS OF WHITE-NOISE MEASURING EQUIPME
F.D.M. MULTI-CHANNEL TELEPHONY

Maximum number of telephone traffic channels	Bandwidth of white-noise spectrum	Centre frequencies of measurement channe				
		Lower channel*		Centre channel	Upper chann	
	kc/s	kc/s	kc/s	kc/s	kc/s	k
60	60–300	50	70	—	270	3
120	60–552	50	70	270	534	6
300	60–1 300	50	70	534	1 248	14
600	60–2 660	50	70	1 248	2 438	32
960	60–4 028	50	70	2 438	3 886	47
	316–4 188	270	534	2 438	3 886	47
1 800	316–8 204	270	534	3 886	8 002	9

* Alternative values.

The detailed characteristics of the white-noise signal and measurement channels have also been defined.

Measurements of the ratio of noise in the freely transm band to that in the measurement channels (the 'noise po ratio) can readily be translated into the corresponding signal/n ratio (the signal being a test tone of 1 mW applied at a zero point and the noise psophometrically weighted), by the version factor given in Fig. 15, subject to some adjustment w

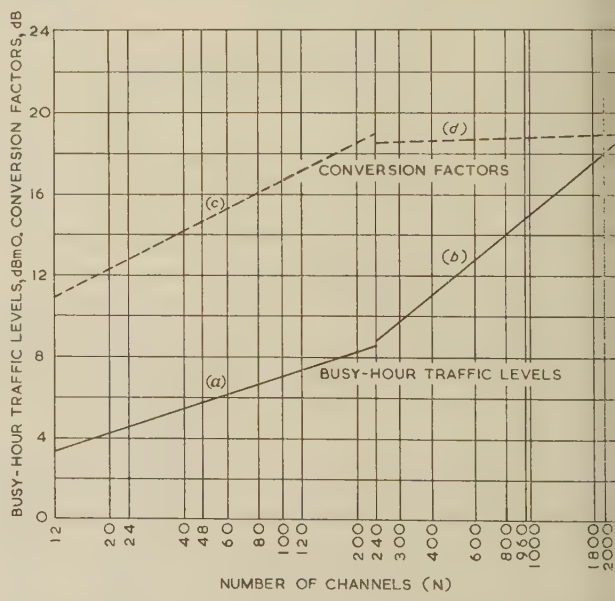


Fig. 15.—Busy-hour traffic levels and conversion factors fo determination of signal/noise ratios from measured noise p ratios.

Busy-hour traffic is expressed as the mean power level of the multi-channel measured at a point of zero relative level.

For $N = 12$ to 240 channels the busy-hour traffic is $(-1 + 4 \log_{10} N)$ curve (*a*).

For $N > 240$ channels, the busy-hour traffic is $(-15 + 10 \log_{10} N)$ dBm0: cu

The conversion factor is added to the measured noise power ratio to gi signal/noise ratio corresponding to 1 mW of tone and the noise power in a 3 band (psophometrically weighted).

The unweighted noise power is 2·5 dB more than the psophometrically we noise power in a 3·1 kc/s band.

the predominating noise contribution arises from a parti type of intermodulation product (Note 3 of Document No. C.C.I.T.T. Study Group No. 1 gives additional informatio

(6.5) Testing Techniques for Television Radio-Relay Syst

The test signals used for the measurement of field-time line-time waveform distortion (test signals Nos. 1 and 2)

on-linearity distortion (test signal No. 3) are described in
x I of Recommendation No. 267.

e low-pass filter used for measurements of random noise is
red to have an insertion loss exceeding 40 dB for frequencies
e $1{\cdot}09f_c$, where f_c is the nominal upper video frequency;
characteristics of the filter are defined in Annex II of
mmendation No. 267.

e weighting network to be used when making measurements
ntinuous random noise has an insertion loss given by

$$10 \log_{10} (1 + \omega^2 T^2) \text{ decibels}$$

e value of T is 0·33 microsec for 405- and 625-line signals,
0·166 microsec for 819-line signals ($f_c = 10$ Mc/s); the
acteristics of the filter are defined in Annex III of Recom-
dation No. 267. The weighting factors for 'white' and
gular' noise are given in Table 8.

Table 8

THEORETICAL NOISE WEIGHTING FACTORS FOR TELEVISION SYSTEMS

System (number of lines and upper video frequency)	Theoretical weighting	
	White noise	Triangular noise
	dB	dB
405 (3 Mc/s)	6·5	12·3
625 (5 Mc/s)	8·5	16·3
625 (6 Mc/s)	9·3	17·8
819 (10 Mc/s)	8·5	16·3

TIME-DIVISION-MULTIPLEX RADIO-RELAY SYSTEMS FOR MULTI-CHANNEL TELEPHONY

ime-division multiplex (t.d.m.) radio-relay systems have so
found only limited application to the international network,
these have been to systems providing relatively small
bers of telephone circuits, say 24. Of the systems in service
majority use pulse-position modulation with amplitude
dulation of the radio carrier (p.p.m.–a.m.). However, p.p.m.
ems with frequency modulation of the radio carrier (p.p.m.–
), and systems using pulse amplitude modulation with
uency modulation of the radio carrier (p.a.m.–f.m.), are
used to a limited extent. Pulse-code modulation (p.c.m.)
e-division multiplex systems do not appear to have been used
civil radio-relay links.

T.D.M. systems have not found wide application in the inter-
ional field for the following reasons:

(i) The international line and radio-relay systems at present in
se employ frequency-division multiplexing almost universally, and
nterconnection with t.d.m. radio-relay links can only be made on
n audio basis (Recommendation No. 304), requiring additional
nodulation and demodulation stages and introducing additional
oise and distortion (the possibility exists, however, of transmitting
number of f.d.m. groups of channels by t.d.m., but it has not been
xploited).

(ii) The radio-frequency bandwidth required for t.d.m. radio-relay
ystems (and particularly for p.c.m. systems) is appreciably wider
han for f.d.m. systems of comparable performance and capacity,
nd this leads to a lack of spectrum economy.

(iii) T.D.M. has not, so far, been shown to be technically satis-
actory for high-capacity radio-relay systems (300 telephone circuits
r more).

The international standardization of t.d.m. radio-relay systems
s not reached a very advanced stage because of their limited
. The main characteristics on which general agreement has
n reached are summarized below. It is recommended that,
ere t.d.m. radio-relay systems are to be employed, the remain-
ing characteristics should be a matter for agreement between the administrations directly concerned (Recommendation No. 299).

(7.1) Preferred Characteristics for T.D.M. Radio-Relay Systems (Report No. 134)

Characteristics of audio channels.—The characteristics of the audio channels should conform with the recommendations of the C.C.I.T.T. for modern telephone circuits and for the method of signalling (Recommendation No. 297).

Maximum number of telephone traffic channels.—To achieve maximum economy in interconnection with other systems, particularly f.d.m. radio-relay systems and line systems, it is highly desirable to provide telephone traffic channels in groups of 12.

Channel sampling rate for telephone traffic.—The preferred channel sampling rate is 8 kc/s with a tolerance of ±8 c/s or better.

Pulse polarity at point of interconnection.—For p.p.m. systems positive polarity is preferred.

Impedance characteristics at point of interconnection.—The preferred nominal impedance at the point of interconnection is 75 ohms.

Channel pulses.—The pulse width and shape should be such as to require the minimum bandwidth consistent with the facilities given by the system.

Grouping of channels using marker-signal synchronization.—If marker signals are used, it is recommended that each group of 12 channels be assigned an individual marker signal to enable the other groups to function properly when one or more groups suffer a breakdown, and to facilitate the branching-off of groups.

Input and output pulse amplitudes.—The preferred pulse amplitude at a point of international interconnection in p.p.m. systems is 1·4 volts at the output from the receiving equipment and 0·7 volt at the input to the transmitting equipment; the difference in level allows for loss in the means of interconnection.

(7.2) Hypothetical Reference Circuits and Circuit Noise in T.D.M. Multi-Channel-Telephony Radio-Relay Systems

It is recommended that the hypothetical reference circuit for t.d.m. radio-relay systems with a capacity of 60 telephone channels or less should be 2 500 km long and comprise six sections of equal length with voice-channel modulation and demodulation at each terminal of a section (Recommendation No. 300).

The circuit noise allowable in the hypothetical reference circuit for t.d.m. radio-relay systems corresponds to that for f.d.m. systems (Section 2.2) with an additional allowance of 2 500 pW for the noise contributed by the t.d.m. terminal equipment (Recommendation No. 301).

(8) TROPOSPHERIC-SCATTER RADIO-RELAY SYSTEMS

Radio-relay systems using the tropospheric-scatter mode of propagation[7, 8] are a comparatively recent development, and the definition of preferred characteristics for such systems has not yet reached a very advanced stage. However, this work is considered urgent, particularly from the aspect of preferred r.f. channel arrangements, since it is clear that international planning and co-ordination of the radio frequencies used is essential in many areas in view of the considerable distances over which the signals may be propagated and the consequent risk of interference both to other tropospheric-scatter systems and to other services sharing the same frequency bands. Attention should also be paid to spectrum economy since multi-section tropospheric-scatter systems tend to require appreciably more frequency space than line-of-sight radio-relay systems of comparable capacity and length. This arises from the long-distance propagation of tropospheric-scatter signals and the need for additional

frequencies to avoid mutual interference between the various sections of the system.

The present position on the international standardization of tropospheric-scatter systems is as follows:

(8.1) Baseband Characteristics

The preferred baseband characteristics for f.d.m. multi-channel telephony, discussed in Section 3.1 and given in Table 2, apply also to tropospheric-scatter radio-relay systems of the appropriate capacity. The adoption of these characteristics will, of course, facilitate interconnection with line-of-sight radio-relay and line systems.

(8.2) Intermediate-Frequency and Modulation Characteristics

No decisions have as yet been made on preferred intermediate-frequency or modulation characteristics for tropospheric-scatter systems, since these may be different from those used for line-of-sight systems in view of the characteristics of tropospheric-scatter propagation (faster fading rate, more marked selective fading and the need to operate with generally lower carrier/noise ratios). Although frequency modulation is generally used for tropospheric-scatter systems, other methods of modulation, including single-sideband amplitude modulation, are under consideration.[19]

(8.3) Radio-Frequency Channel Arrangements

As already noted, the preparation of preferred r.f. channel arrangements for tropospheric-scatter systems is a matter of some difficulty owing to the many frequencies needed to avoid interference between the sections of multi-section systems arising from the long-distance propagation of tropospheric-scatter signals. It is further complicated by the need to avoid interference between the high-power transmitters and the low-noise receivers at a given station, which may, for reasons of economy, have to share the same aerial. The r.f. channel arrangements adopted must also have considerable flexibility in order to accommodate systems with different telephone-channel capacities and perhaps one or more television channels. At the time of the IXth Plenary Meeting of the C.C.I.R. (Los Angeles, 1959) no frequency bands had been allocated internationally to tropospheric-scatter systems, and it was not then known how much frequency space, if any, would be available for such systems. The allocations subsequently made by the Administrative Radio Conference (Geneva, 1959) are indicated in Section 8.5. The best that could be done, therefore, at the IXth Plenary Meeting of the C.C.I.R. was to prepare a Report on the technical basis for preparing r.f. channel arrangements as a guide to administrations (Recommendation No. 303 and Report No. 136). The r.f. channel arrangements shown in the Report will need revision in the light of the frequency allocations subsequently made by the Administrative Radio Conference.

(8.4) Limitation of Interference

The C.C.I.R. has noted that tropospheric-scatter systems require a high degree of international co-ordination and planning to avoid interference between systems in nearby countries, and that the problem becomes much more complex if such systems occupy the same frequency bands as conventional line-of-sight systems or other services (Recommendation No. 302). In stressing the need for the utmost frequency economy it recommends that frequency diversity should be avoided, as far as possible, particularly in those parts of the world where the spectrum is likely to become congested. To reduce interference special efforts should be made to operate tropospheric-scatter systems at the lowest practicable level of radiated power, to reduce radiation in and reception from undesired directions an reduce spurious emissions to the lowest practicable level.

(8.5) Frequency Bands for Tropospheric-Scatter Radio-R Systems

The frequency allocations made by the Administrative R Conference (Geneva, 1959) for tropospheric-scatter radio- systems, and the conditions under which these allocations be used, are given in Footnotes 333 and 364 of the Frequ Allocation Table (Article 5 of the Radio Regulations):

> 333. In Region I, stations of the fixed service using troposp sca‹'er may operate in the band 790–960 Mc/s subject to a ments between the administrations concerned and affected. operations in the band 790–860 Mc/s shall be on a secondary to those of the broadcasting service.
>
> 364. In Region I, tropospheric-scatter systems may oper the band 2 550–2 700 Mc/s under agreements concluded be administrations concerned and those having services operati accordance with the Table, which may be affected.

(8.6) Hypothetical Reference Circuit and Circuit Noise Tropospheric-Scatter Systems (Report No. 135)

No formal recommendation for the hypothetical refe circuit for tropospheric-scatter systems has as yet been m but an overall length of 2 500 km has been proposed. It been noted that it is not at present possible to propose a division into homogeneous sections since the length of systems may vary widely, e.g. between 100 and 400 km. further noted that the interconnection between systems usually be made at baseband frequencies.

The question of circuit noise is currently being studied b C.C.I.T.T./C.C.I.R. Joint Working Party on Noise. It is re mended by the C.C.I.R. that, so far as mean noise pow concerned, the value of 7 500 pW mean psophometric pow any hour at the end of a 2 500 km hypothetical reference ci (Section 2.2) is achievable on tropospheric-scatter systems that it is important to maintain such performance standard international circuits. However, this desirable objective m some cases, in the present state of the art, result in prohib or excessive cost, impracticably high transmitter power or power as might result in undesirable interference. The app tion of the stated noise objective might also make impracti some desirable extensions of the telephone network. A l standard of performance, e.g. similar to that adopted by C.C.I.T.T. for open-wire lines, is envisaged for such cases.

(9) FUTURE DEVELOPMENTS

As noted earlier the design of microwave radio-relay sys is not static, and developments will undoubtedly necessitat revision and extension of the preferred characteristics.

Although most of the international microwave radio- systems now in use employ frequency modulation of the r carrier, it is possible that increasing demands on spectrum s and the need for still greater frequency economy will ind the desirability of using other forms of modulation suc single-sideband amplitude modulation. The latter would the prospect of a considerable saving in bandwidth, alth at the expense of more stringent linearity requirements and additional complexity of the equipment required.

The maximum capacity at present envisaged on micro links is 1 800 telephone channels, or the equivalent, on each carrier. A higher capacity with, for example, the same cap as the C.C.I.T.T. 12 Mc/s coaxial-cable system (2 700 telep channels or the equivalent) would appear to be the next in development. However, more information is needed the multi-path transmission and selective fading character

pical microwave paths before it can be said that such ns are practicable. This is the subject of a C.C.I.R. Programme.

ere is scope for the further development of the means for mitting a number of high-quality music channels on the radio carrier as a television channel, e.g. by the use of f.m. arriers. Such a development would be valuable for the mission of composite sound and vision signals over inter- nal circuits such as the Eurovision network.

e present preferred characteristics of line-of-sight micro- systems are primarily concerned with systems operating at encies up to 8·5 Gc/s. A number of frequency bands e 8·5 Gc/s have been allocated by the Administrative Radio erence (Geneva, 1959) for use by fixed systems, and some ese may well find application to the provision of inter- nal circuits for television and multi-channel telephony, bly using shorter paths than is customary at lower radio encies.

the case of tropospheric-scatter systems further studies are ed of the performance achievable, the maximum transmis- capacity that is practicable, the modulation methods and acteristics to be used, and the r.f. channel arrangements to mployed. The latter study is urgent in view of the long- e interference aspects of tropospheric-scatter systems and possible repercussions on international usage of the radio trum.

o mention has so far been made of microwave systems g earth satellites, but it seems clear that such systems are ined to play an important part in world-wide communica- and that international planning and co-ordination will be ntial. This task has been entrusted to the newly-formed dy Group IV (Space Systems) of the C.C.I.R.

(10) ACKNOWLEDGMENTS

hanks are especially due to Dr. E. Metzler, Director of C.C.I.R., for kind permission to make extensive use of the uments and work of the C.C.I.R. in the present paper; onsibility for the use made in the paper must, however, be olly that of the author.

he work described is that of many engineers in many countries was carried out under the auspices of Study Group IX of the .I.R., of which the author is privileged to be the present rnational Chairman. He would like to take this opportunity hanking the Vice-Chairman, Mr. E. Dietrich, and the members Study Group IX for their willing co-operation and help. ticularly he would like to acknowledge the valuable founda- n to this work laid by the immediate Past Chairman of Study oup IX, Mr. H. Stanesby.

Tribute is also due to the engineers of the Bell Telephone oratories, whose pioneering work on microwave radio-relay tems has greatly helped such systems to take their rightful ce in the ever-growing family of international and national nsmission systems.

Acknowledgment is also made to the Engineer-in-Chief of the neral Post Office for permission to make use of the informa- n contained in the paper.

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(11.2) Reference to Recommendations and Reports of the C.C.I.R.

The following Recommendations and Reports are from the documents of the IXth Plenary Assembly of the C.C.I.R. (Los Angeles, 1959).

Recommendations.

No. 267. Requirements for the Transmission of Monochrome Television Signals over Long Distances.

No. 268. Radio-Relay Systems for Telephony using Frequency-Division Multiplex. (Interconnection at Audio Frequencies.)

No. 269. Radio-Relay Systems for Telephony using Frequency-Division Multiplex. (Interconnection at Baseband Frequencies.)

No. 270. Radio-Relay Systems for Television. (Interconnection at Video Signal Frequencies.)

No. 271. Radio-Relay Systems for Television and Telephony. (Simultaneous Transmission by the same Radio Frequency Carrier. Baseband Arrangements.)

No. 272. Radio-Relay Systems for Television. (Simultaneous Transmission of the Monochrome Television Signals and a Single Sound Channel. Preferred Characteristics of the Sound Channel.)

No. 273. Radio-Relay Systems for Television and Telephony. (Intermediate Frequency Characteristics.)

No. 274. Radio-Relay Systems for Telephony using Frequency-Division Multiplex. (Frequency Deviation.)

No. 275. Radio-Relay Systems for Telephony using Frequency-Division Multiplex. (Pre-Emphasis Characteristics for Frequency-Modulation Systems.)

No. 276. Radio-Relay Systems for Television. (Frequency Deviation and the Sense of Modulation.)

No. 277. Radio-Relay Systems for Television. (Pre-emphasis Characteristics for Frequency-Modulation Systems.)

No. 278. Radio-Relay Systems for Television and Telephony. (Radio-Frequency Channel Arrangements for Systems for 600 to 1 800 Telephone Channels, or the Equivalent, Operating in the 2 000 and 4 000 Mc/s Bands.)

No. 279. Radio-Relay Systems for Telephony using Frequency-Division Multiplex. (Radio-Frequency Channel Arrangements for 300 Channel Systems Operating in the 2 000 and 4 000 Mc/s Bands.)

No. 280. Radio-Relay Systems for Television and Telephony. (Radio-Frequency Channel Arrangements for Systems for 600 to 1 800 Telephone Channels, or the Equivalent, Operating in the 6 000 Mc/s Band.)

No. 281. Radio-Relay Systems for Television and Telephony. (Preferred Radio-Frequency Channel Arrangements for Television.)

No. 282. Radio-Relay Systems for Television and Telephony. (Use of Special Radio-Frequency Channel Arrangements.)

No. 283. Radio-Relay Systems for Telephony using Frequency-Division Multiplex. (Radio-Frequency Channel Arrangements for 60- and 120-Channel Telephony Systems Operating in the 2 000 Mc/s Band.)

No. 284. Radio-Relay Systems for Telephony using Frequency Division Multiplex. (Radio-Frequency Channel Arrangements for 60- and 120-Channel Telephony Systems Operating in the 7 000 Mc/s Band.)

No. 285. Radio-Relay Systems for Telephony using Frequency-Division Multiplex. (Hypothetical Reference Circuit for Radio-Relay Systems with Capacity of 12 to 120 Telephone Channels.)

No. 286. Radio-Relay Systems for Telephony using Frequency-Division Multiplex. (Hypothetical Reference Circuit for Radio-Relay Systems with Capacity of more than 120 Telephone Channels.)

No. 287. Radio-Relay Systems for Telephony using Frequency-Division Multiplex. (Allowable Noise Power in the Hypothetical Reference Circuit.)

No. 288. Radio-Relay Systems for Telephony using Frequency-Division Multiplex. (Noise in Real Circuits.)

No. 289. Radio-Relay Systems for Monochrome Television. (Permissible Noise in the Hypothetical Reference Circuit.)

No. 290. Maintenance Procedure for Radio-Relay Systems for Telephony using Frequency-Division Multiplex. (Measurements to be made.)

No. 291. Interconnection of Radio-Relay and Line Systems. (Line-Regulating and other Pilots.)

No. 292. Radio-Relay Systems for Television and Teleph (Frequencies and Deviations of Continuity Pilots.)

No. 293. Radio-Relay Systems for Telephony using Freque Division Multiplex. (Maintenance Measurements in A Traffic.)

No. 294. Radio-Relay Systems for Telephony using Freque Division Multiplex. (Measurements of the Performance the Help of a Signal consisting of a Continuous Uni Spectrum.)

No. 295. Service Channels for Radio-Relay Systems. (Ty Service Channels to be Provided.)

No. 296. Radio-Relay Systems for Television and Teleph (Preferred Characteristics of Auxiliary Radio-Relay Sys Operating in the 2 000, 4 000 or 6 000 Mc/s Bands.)

No. 297. Radio-Relay Systems for Telephony using T Division Multiplex. (Interconnection at Audio Frequen

No. 298. Radio-Relay Systems for Telephony using T Division Multiplex. (Preferred Characteristics.)

No. 299. Radio-Relay Systems for Telephony using T Division Multiplex. (Agreement on Major Characteris

No. 300. Radio-Relay Systems for Telephony using T Division Multiplex. (Hypothetical Reference Circui Radio-Relay Systems with Capacity of 60 Telephony Ch or less.)

No. 301. Radio-Relay Systems for Telephony using T Division Multiplex. (Allowable Noise Power in the H thetical Reference Circuit.)

No. 302. Radio-Relay Systems Employing Troposp Scatter Propagation. (Limitation of Interference.)

No. 303. Radio-Relay Systems using Tropospheric-S Propagation. (Radio-Frequency Channel Arrangements.

No. 304. Radio-Relay Systems for Telephony. (Interco tion of Differing Multiplex Systems.)

No. 305. Radio-Relay Systems for Television and Teleph (Standby Arrangements.)

No. 306. Radio-Relay Systems for Television and Teleph (Procedure for International Connection between Systems Differing Characteristics.)

Reports.

No. 129. Radio-Relay Systems for Telephony using Freque Division Multiplex. (Calculation of Intermodulation N due to Non-Linearity.)

No. 130. Radio-Relay Systems for Telephony. (Noise T able during very Short Periods for Time on Line-of-S Systems.)

No. 131. Radio-Relay Systems for Telephony using Freque Division Multiplex. (Technical Characteristics to be Spec in order to Connect any Two Systems.)

No. 132. Radio-Relay Systems for Telephony using Freque Division Multiplex. (Design Objectives for Voice-Frequ Telegraphy on Telephone Channels.)

No. 133. Radio-Relay Systems for Television and Teleph (Alternative Transmission of Television and Telephony.)

No. 134. Radio-Relay Systems for Telephony using T Division Multiplex. (Technical Characteristics to be Spec in order to Connect any Two Systems.)

No. 135. Radio-Relay Systems using Tropospheric or I spheric Forward Scatter.

No. 136. Radio-Relay Systems using Tropospheric-Sc Propagation.

No. 137. Duration of Interruptions on Radio Links Switching from Normal to Standby Equipment.

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R.F. SWITCHING CIRCUITS AND HYBRID RING CIRCUITS USED IN RADIO ASTRONOMY

By F. G. SMITH, M.A., Ph.D.

(*The paper was first received* 1*st July, and in revised form* 8*th October*, 1960).

SUMMARY

e use of r.f. switching and hybrid ring circuits in radio astronomy s from the necessity of comparing small r.f. noise powers without ducing lossy switching elements in the receiver input. At fre- cies where coaxial techniques are appropriate r.f. switches nonly use crystal diodes. This paper compares the series and connections of such switches, showing that the minimum loss e same but that the practical details usually favour the stub ection.

r the half-wave switching function used in interferometer aerial ms, the stub diode switch is shown to be a special case of the d ring. Hybrid rings may be used either as splitter circuits or idges for comparing impedances. Reference is made to these cations in radio astronomy.

(1) INTRODUCTION

eceivers used with practically all types of radio telescope n a high degree of sensitivity and stability by recording r a difference between two signal sources[1] or a correlation een them.[2] Both techniques avoid the direct recording of ver noise, and both can use a variety of switching circuits hybrid circuits. The paper is concerned with the range of circuits useful in receivers for frequencies where coaxial niques are appropriate; the main difference in waveguide its arises from the existence of non-reciprocal components, h are not considered here.

he switching circuits are either change-over switches, which be single or double pole, or they are phase-changing ches introducing an extra electrical path in a coaxial circuit. tal diodes have been found to be the most suitable switching ents. It is often important to choose the switching circuit h introduces the least loss, as, for example, when the switch be used in a radiometer using a receiver with a very low t noise temperature. The paper therefore contains an ysis of the loss in the various switching circuits and shows any particular diode may be used in such a way that the is a minimum.

he hybrid circuits are used in splitting the output of a rator between two receivers, and also as a form of bridge uit to compare impedances. A combination of these func- s has been used for the absolute calibration of noise signal rators against thermal loads. A brief analysis of the circuits onstrates their main properties; it may be used, for example, n analysis of the accuracy of a bridge circuit using a hybrid .

(2) SWITCHES

(2.1) Switching Diode

appears that the main parameter affecting the loss and ation in r.f. switches is the ratio of the admittances of the nary switching element in its 'on' and 'off' conditions. tching diodes such as the T19G and the OA47 may, for our purposes, be considered as a fixed resistance R in series with a capacitance which changes from a low value C to a very high value whose impedance is much less than R. The admittance therefore changes from $1/R$ to $R\omega^2C^2 + j\omega C$. Since C may usually be tuned out, the switching ratio K is basically $1/R^2\omega^2C^2$. This ratio may be written ω_0^2/ω^2, where ω_0 is a high-frequency characteristic of the diode. For example, a diode with $R = 10$ ohms and $C = 1$ pF, $\omega_0/2\pi = 1{\cdot}6 \times 10$ Gc/s; such a diode would allow effective switching up to frequencies of several gigacycles per second.

(2.2) Series Single-Pole Change-Over Switch

Fig. 1 represents a load resistance R_L, such as the input of a receiver, connected to either of two generators according to the

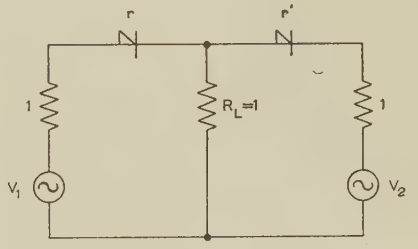


Fig. 1.—Series change-over switch.

d.c. bias on the diodes. The diodes are now considered to have resistive impedances r and r', where

$$r' = \frac{1}{r\omega^2C^2}$$

The d.c. biasing arrangements and the tuning for C are not shown. The generators are matched to the receiver input, and all impedances are normalized to the generator impedance. By a suitable choice of this impedance, or by a suitable transformation in the transmission lines on either side of each diode, it is possible to make $rr' = 1$; if the ratio $K = r'/r$ is large, this will give the least loss. In this condition the transmitted power is $1 - (2/\sqrt{K})$ of the optimum. The ratio of the powers received from the 'on' and 'off' sides is then approximately equal to K; a larger ratio can be obtained at the expense of increasing the loss in the switch.

A difficulty in realizing this performance arises in the practical arrangement for tuning out C, since an inductance must be placed in parallel with the diode either as a lumped impedance or as a section of balanced stub line. We shall find that the same switching performance can be obtained for a stub switch, with greater convenience.

(2.3) Stub Single-Pole Change-Over Switch

In Fig. 2 the crystal diodes are connected to the switching points B and D via quarter-wave transformers with impedance Z_T, which may be chosen to give the least loss in the switch.

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. Smith is at the Mullard Radio Astronomy Observatory.

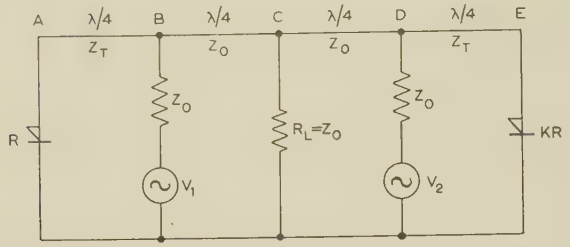


Fig. 2.—Stub change-over switch.

The diodes appear at B and D as admittances R/Z_T^2 and KR/Z_T^2, which may be considered, respectively, small and large compared with Z_0, the impedance of the generators, the load, and the lines BC and DC.

The power transferred from the generator V_1 to the load R_L is attenuated by the factor

$$1 - \frac{Z_T^2}{KRZ_0} - \frac{RZ_0}{Z_T^2}$$

which again reduces to the best value $[1 - (2/\sqrt{K})]$ when $Z_0^2 = Z_T^4/KR^2$. Since $KR^2 = 1/\omega^2C^2$ we find that the best value of Z_T is $(Z_0/\omega C)^{1/2}$.

This condition does not give the largest switching ratio (the ratio of powers received from the two generators). Since it is, in practice, impossible at low frequencies to obtain the theoretical loss of $(1 - 2\omega/\omega_0)$, which may easily be less than 0·1 dB, it may be considered desirable to use a value of Z_T rather lower than $(Z_0/\omega C)^{1/2}$. The switching ratio may then be considerably improved without appreciably increasing the loss.

The practical advantage of this form of switch over the series switch lies in the ease of tuning out C, which may be achieved by shortening the transformers BC and CD slightly, and adding a small capacitance at B and D.

A stub switch of this kind has been made for a frequency of 408 Mc/s using T19G diodes. The measured loss was 0·25 dB, compared with a theoretical loss of 0·20 dB. Diodes which have been developed for parametric amplifiers, as for example the VX3314, have higher cut-off frequencies. The loss in a stub switch at 408 Mc/s using two VX3314 diodes would be 0·1 dB.

(2.4) Series λ/2 Switch

The single-pole switch may be adapted to switches which insert extra lengths of line into a transmission-line circuit. The insertion of a half wavelength, as in Fig. 3, is required in phase-

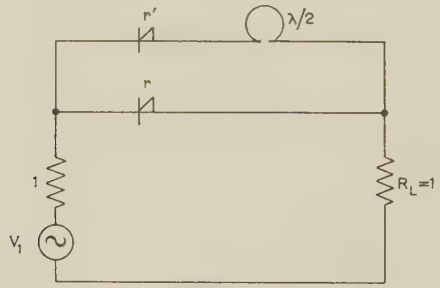


Fig. 3.—Half-wave switch.

switching interferometers, in which two aerials are joined to one receiver alternately in and out of phase.

Maximum power is delivered by this circuit when $rr' = 4$. The phase reverses exactly when r and r' interchange. Again it is difficult to tune out C, or to transform the impedances r and r' to obtain the least loss. With no attempt at e improvement, this switch will generally only be useful at quencies below about 300 Mc/s.

(2.5) Series Switches for Other Phase Changes

When lengths other than $\lambda/2$ are to be inserted, it is nece to use two change-over switches, as in Fig. 4; otherwise

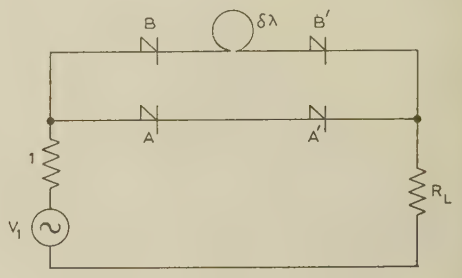


Fig. 4.—Insertion of extra cable length $\delta\lambda$.

would be a large mismatch in one position of the switch. the signal passes via the diodes AA′ or BB′, according d.c. bias on them.

A complete equality of mismatch between the two posi of the switch is obtained only by inserting extra lengths of in both paths; for the quarter-wave switch the two path and BB′ would then be $\lambda/8$ and $3\lambda/8$, respectively.

As before, if the least possible loss is required, it ma necessary to use stub switches instead of series at both en this insertion switch. For the particular case of a $\lambda/2$ inser the circuit becomes a form of hybrid ring, as considered i next Section.

(3) HYBRID RINGS IN SWITCHING

(3.1) Phase-Switching Hybrid Ring (see Fig. 5)

The phase-switching hybrid-ring circuit may be regarde two stub switches at the points A and C, or, more properly, particular case of the hybrid ring. The more general u

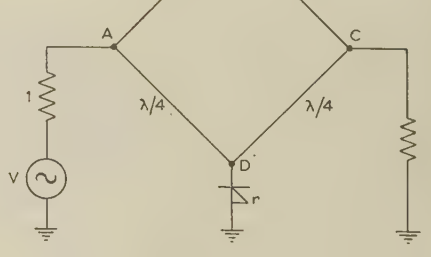
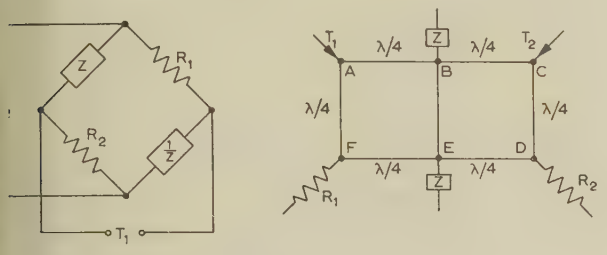


Fig. 5.—Half-wave hybrid switch.

hybrid rings is considered later; here we note that, if the i dances at B and D are interchanged, the output at C rev phase exactly. Again the normalized impedances r and r' be obtained by transforming the impedances R and KR' thr $\lambda/4$ transformers to suit the value of Z_0 of the circuit. The power transfer is $(1 - 4/\sqrt{K})$.

(3.2) Double-Pole Change-Over Switch

is occasionally advantageous to use two receivers when a parison is made between two sources of noise power, ing the connection by means of a change-over switch. problem is closely related to that of duplexing two trans- ers on to a single aerial, which is commonly achieved by circuit of Fig. 6(*a*).



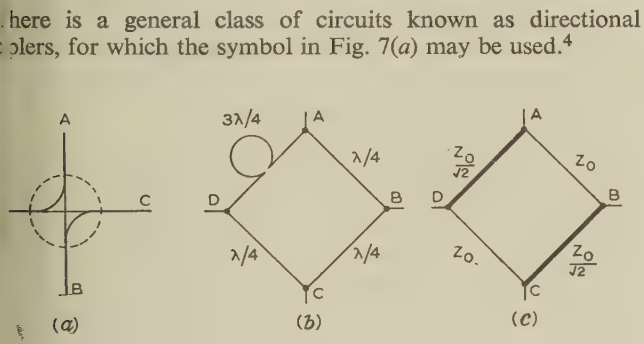
(*a*) (*b*)

Fig. 6.—Duplexing networks.

this bridge circuit the loads R_1 and R_2 have the same edance R_0. At frequencies where Z is low T_1 is connected $_1$ and T_2 to R_2; when Z becomes high the connections are changed. It has been shown by Phillips[3] that the coaxial uit of Fig. 6(*b*) gives an equivalent performance when all ths are quarter wavelengths. This is clearly adaptable to our pose, by connecting switching diodes at B and E. The use of a double-pole change-over switch in a measurement of noise er is described in Section 5.2.

) HYBRID CIRCUITS AND TRANSFORMER HYBRIDS

here is a general class of circuits known as directional plers, for which the symbol in Fig. 7(*a*) may be used.[4]



(*a*) (*b*) (*c*)

Fig. 7

(*a*) Directional coupler.
(*b*) 180° hybrid.
(*c*) 90° hybrid.

onnecting lines in this symbol indicate, for example, that a erator at A would transmit power to B and D but not to C, vided that all terminals were matched. The ratio of the puts at B and D may be very high; this is the usual arrange- t when a directional coupler is used to measure the reflection fficient of the load at B by comparing outputs at C and D. ernatively, if the ratio of the outputs at B and D is arranged e unity, the circuit has several uses as a splitter and as a lge. We shall review various practical forms of such a rid.

(4.1) 180° Hybrid

he hybrid of Fig. 7(*b*) is made of coaxial cable with the same racteristic impedance Z_0 throughout. Normalizing all impe- ces and admittances to this value, it may be shown that if a rent generator I_A with output admittance G_A is connected A, the following currents flow in the admittances G_B, G_C G_D connected between B, C, D and earth:

$$I_B = -\frac{j}{\Delta}(G_C G_D + 2)\frac{G_B}{G_A} I_A$$

$$I_D = \frac{j}{\Delta}(G_B G_C + 2)\frac{G_D}{G_A} I_A$$

$$I_C = \frac{G_D - G_B}{\Delta}\frac{G_C}{G_A} I_A$$

where $$\Delta = G_A G_B G_C G_D + \sum_{x,y} G_x G_y + 4$$

The main features of this circuit at its centre frequency are as follows:

(i) The points A and C are isolated from one another if equal admittances are presented at B and D.
(ii) The current at C reverses in phase if the admittances at B and D interchange (this is the action of the switch in Section 3.1).
(iii) When admittances $\sqrt{2}G_0$ (where $G_0 = 1/Z_0$) are presented at B and D, a generator at A with output admittance $\sqrt{2}G_0$ is matched, and equal anti-phased currents flow at B and D.

(4.2) 90° Hybrid

In Fig. 7(*c*) all four arms are quarter wavelengths, but AB and CD have characteristic impedance Z_0, and AD and BC have characteristic impedance $Z_0/\sqrt{2}$. This circuit behaves like the 180° hybrid except that

(i) A and C are isolated if $G_B = 1/G_D$.
(ii) The hybrid is matched by impedances Z_0 when the currents at B and D are in quadrature.

There appears to be no advantage in this over the hybrid circuit shown in Fig. 7(*b*), which has been more widely used in radio astronomy.

(4.3) Transformer Hybrid

The transmission-line hybrids are strictly correct at only one frequency. At lower frequencies where lumped circuits are practical the transformer bridge of Fig. 8 provides a similar function without frequency sensitivity.

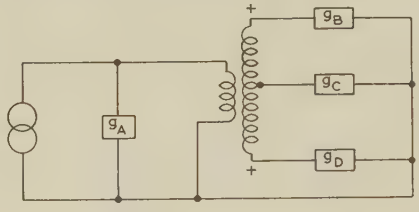


Fig. 8.—Transformer hybrid.

Here the transformer has two equal secondary windings in anti-phase. The use of this circuit as the transformer ratio-arm bridge is well known, but there is also a close correspondence with the action of the hybrid circuits in their properties of isolation and splitting.

(5) MEASUREMENT OF SMALL R.F. NOISE POWERS

Comparison of the noise powers in two generators, as, for example, in an aerial and a diode noise source, has been made by using receivers with simple change-over switches at the input, and also by using phase-switching receivers. The two methods are basically the same, and both suffer from the disadvantage that the receiver output is a function of the impedance of the source as well as of its available noise power. This arises from the dependence of the receiver noise on the source impedance, and errors of noise temperature corresponding to appreciable fractions of the receiver noise temperature may occur. It appears that, in any such measurement, it is necessary to be able to check the equality of the source impedances, and when low temperatures are involved this check must be very accurate.

Two methods, adaptable to the two kinds of receiver, are described first. In Section 5.3 a receiver which is comparatively insensitive to impedance changes is described.

(5.1) Phase-Switching Receiver with a Hybrid Ring

In the arrangement of Fig. 9, the two amplifiers 1 and 2 ideally receive noise signals $V_{N1} - V_{N2}$ and $V_{N1} + V_{N2}$, respectively, from the 180° hybrid ring ABCD. The receiver output, after these have been added via a commutating 180° switch and after simple square-law detection, is proportional to $V_{N1}^2 - V_{N2}^2$. The required check of impedances may be made by disconnecting amplifier 2 from the point D, and feeding in an r.f. signal at this point from a high-level signal generator. Any inequality of impedances will allow some of this signal to enter the receiver via A_2, so that a matching network M may now be adjusted for zero transmission.

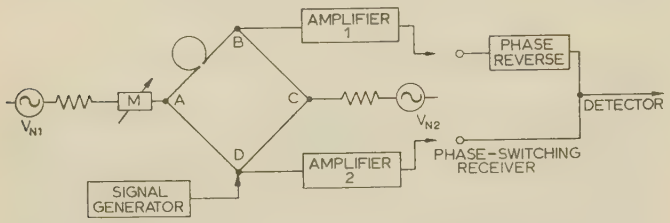


Fig. 9.—Comparison of noise powers using a phase-switching receiver.

This method has been used by Costain[5] for accurate measurement of equivalent aerial temperature in the metre wavelength range.

(5.2) Simple Change-Over Switch, with Directional Coupler

The incorporation of a directional coupler into a switched receiver, as in Fig. 10, allows the match of either load to be checked at any time by stopping the switch and injecting power from the high-level signal generator. If the loads on B and D are equal, no power will appear in the receiver. A possible procedure in a comparison of noise powers would be as follows, which would allow for changes in impedance of the source V_{N1} to be removed by the matcher M_1:

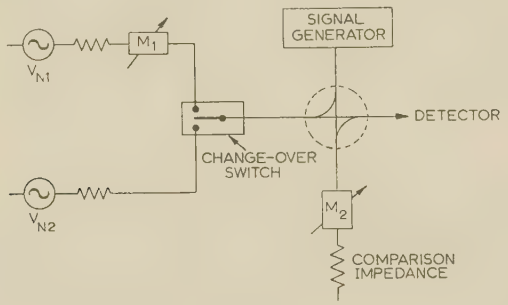


Fig. 10.—Comparison of noise powers using a Dicke receiver.

(i) Stop the switch on the side of V_{N2}.
(ii) Adjust the matcher M_2 for balance.
(iii) With the switch running, adjust M_1 so that there is no change in detected receiver output when the high-level generator is keyed 'on' and 'off'.

This procedure enables the check on impedances to be carried out during the measurement itself.

The directional coupler must usually have a high output ratio, otherwise it will introduce attenuation between the sources and the receiver.

(5.3) Double-Pole Change-Over Receiver

The arrangement of Fig. 11, using for example the switch Fig. 6(*b*), at first sight gives only a slight advantage over simple switch, in that the random noise level of the differ

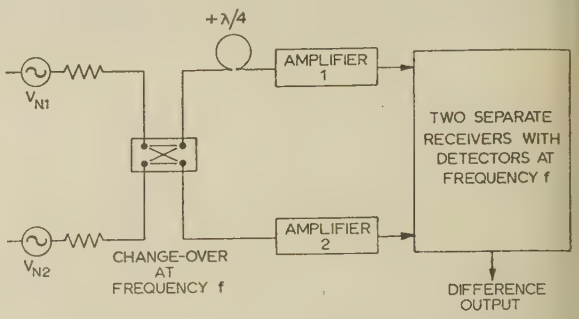


Fig. 11.—Comparison of noise powers using a double-pole double-throw switch.

output is $1\frac{1}{2}$ dB down on that of a single receiver. Howeve the two pre-amplifiers 1 and 2 have identical input cha teristics, the inclusion of an extra quarter wavelength of l one amplifier input reverses the effect of a small mismat that amplifier, so that, as the amplifiers are connected alter to each noise source, the effect on the switch-frequency c ponent of the output is the same in each. The output is obt from the difference in outputs, so that a first-order cancel of the effect of impedance changes is possible.

(6) CONCLUSIONS

The uses which have been made of these circuits in r astronomy may be summarized as follows:

The series change-over switch of Section 2.2 and the s half-wave switch of Section 2.4 have been widely used at lower frequencies and where it is not essential to reduce to a minimum.

The corresponding stub switches of Section 2.3 and 3.1 be used where low loss is important, and especially at frequencies.

Hybrid rings have been used for splitting the signal fro aerial between two receivers when it is required to isolate receivers from one another. For example, an aerial ma connected to a receiver which is measuring the total p received, and at the same time to another receiver which c bines the signal with that from another aerial, formin interferometer system.

Hybrid rings, and the related directional couplers, have used in measurements of small noise powers in which essential to compare the impedances of two sources of power. For example, the noise power from an aerial rece cosmic radio noise can be compared with the power fr heated resistor. This has been accomplished by using circuit arrangements of Section 5.

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A COAXIAL CONNECTOR SYSTEM FOR PRECISION R.F. MEASURING INSTRUMENTS AND STANDARDS

By D. WOODS, Associate Member.

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SUMMARY

neral design requirements are discussed for a coaxial connector m for precision r.f. impedance and admittance standards, bridges measuring instruments of all types for frequencies up to 4 Gc/s. description is given of a practical connector system which has in use for many years at a Ministry of Aviation laboratory. The n is such that the connector system does not introduce uncer- es greater than about 3 parts in 10^4 in the admittance parameter. e residual parameters of the connector system, including the -circuit and short-circuit conditions, are described and evaluated. relating to the discontinuity capacitances encountered in a ical measuring system are also given, together with formulae he precise transformation of admittance between two planes in form coaxial line.

LIST OF SYMBOLS

b = Radii of inner and outer conductors of a coaxial line, cm.
c = Displacement of the axis of a coaxial line due to eccentricity.
$\beta = \omega/v$, rad/cm.
C = Total static capacitance of a coaxial line.
C' = Effective total capacitance of a coaxial line $= (C \tan \beta l)/\beta l$.
C_D = Discontinuity capacitance of a coaxial line with a discontinuous inner conductor.
C_d = Discontinuity capacitance between coaxial lines of different characteristic impedances.
C_f = Terminal fringe capacitance.
C_i = Component of Y_i.
C_r = Component of Y_r.
$\Delta C = C_D - C_f$.
C_t = Terminal capacitance of an immittance meter.
ϵ = Permittivity.
ε = Base of Naperian logarithms.
f = Frequency.
G_f = Radiation conductance at open end of a coaxial line.
G_i = Component of Y_i.
G_r = Component of Y_r.
l = Length, cm.
L_∞ = Limiting value of inductance of coaxial line at zero skin depth.
L = Inductance of a coaxial line (finite skin depth) $= L_\infty(1 + \zeta)^2$.
R_t = Terminal resistance of an immittance meter.
ρ = Resistivity, ohm-cm.
v_∞ = Velocity of electromagnetic waves in air at 20° C, 760 mm Hg and 50% r.h. $= 2{\cdot}9970 \times 10^{10}$ cm/s (i.e. the limiting velocity in a coaxial system when the skin depth is zero).
v = Velocity of propagation in a coaxial system at 20° C, 760 mm Hg and 50% r.h. and for finite skin depth $= v_\infty/(1 + \zeta)$.
$\omega = 2\pi f$.
$Y_i = G_i + j\omega C_i$ = Input admittance of coaxial line terminated by Y_r.
Y_m = Admittance measured by bridge and uncorrected for C_f.
Y_r = Terminating admittance of coaxial line $= G_r + j\omega C_r$.
Z_c = Characteristic impedance of the connector line.
Z_f = Impedance of short-circuit $= R_f + j\omega L_f$.
Z_∞ = Limiting value of characteristic impedance for zero skin depth.
Z = Characteristic impedance of a coaxial line (finite skin depth) $= Z_\infty(1 + \zeta)$.
$1 + \zeta$ = Reciprocal velocity factor $= v_\infty/v$.

(1) INTRODUCTION

When the design of a precision dual admittance bridge[1] was under review it was apparent that plugs and sockets in current use would be unsuitable for precise measurements owing to, first, the discontinuities they produce, and secondly, the inherent uncertainty in the position of the plane of cleavage between mating connectors. The existence of numerous types of plug and socket also leads to difficulty and inconvenience in the design of adaptors to connect one system to another. The almost universal use of male and female terminations is a further disadvantage when two plugs or two sockets have to be connected together. This is usually accomplished by means of a length of flexible coaxial cable fitted with terminations of the same type. The characteristics of flexible cables are not sufficiently stable to allow their use in a precision measuring system where immittance* is required to be known to better than a few per cent. In addition, single-screened types usually leak and the connection of the cable to its terminations introduces large discontinuities.

The requirements, therefore, demanded a connector system of new design such that the effect of a discontinuity at a junction was either negligible or calculable to the required precision in comparison with the overall accuracy required, namely about 0·1%. It was decided to make the system as universal as possible and suitable for r.f. measuring instruments of all types and for all frequencies up to 4 Gc/s, which is the highest frequency at which coaxial systems are normally used. The hermaphrodite principle was adopted as far as possible in order to achieve maximum flexibility in use.

(2) GENERAL DESCRIPTION

The system comprises the following four types of termination:

Type	*Use*
B2	Immittance standards.
B1	Immittance meters.
A1	All other types of measuring instruments.
A2	Simplified version of A1 for use where immittance is not a major parameter.

* *Immittance* is used to convey the general idea of impedance or admittance.

ritten contributions on papers published without being read at meetings are ed for consideration with a view to publication.
r. Woods is at the Ministry of Aviation.

The design is such that the following types can be directly connected together: B2–B1, B1–A1, A1–A1, A1–A2. The need for other combinations does not normally arise.

In addition, there is a coaxial air-spaced connector line fitted with an A1 termination at each end which performs the functions of flexible coaxial cables in conventional measuring systems; it can also be used to connect together terminations of the following types when this is required for a special purpose: B1–A2, A2–A2.

Outline drawings of the connectors, with principle dimensions, are shown in Fig. 1. It will be seen that the inside diameter of

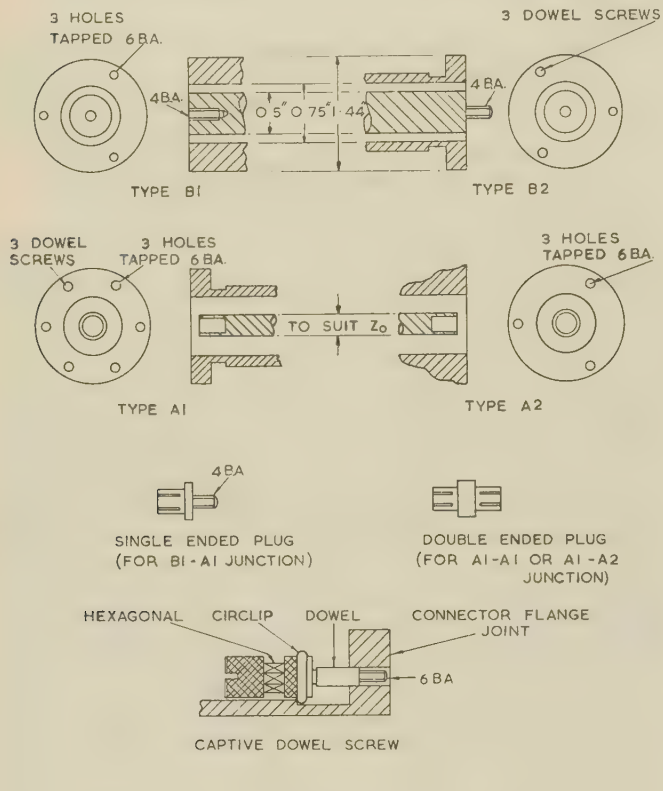


Fig. 1.—General design of connectors.

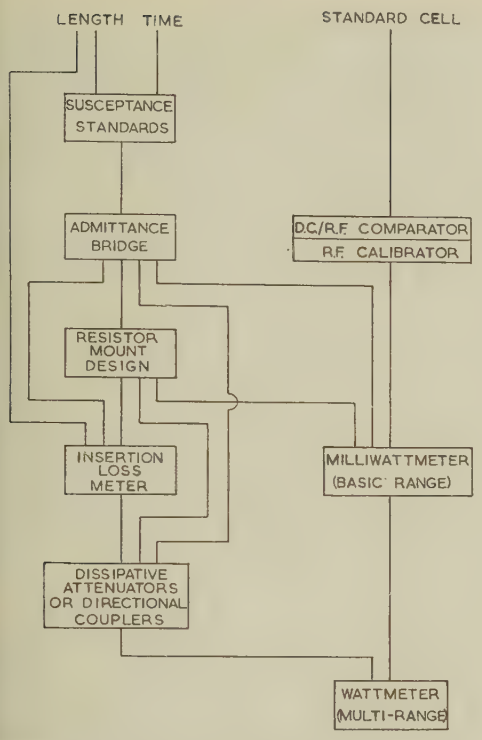
the outer conductor has been standardized at 0·750 in while the outer diameter of the inner conductor is made to suit the characteristic impedance required, except in the case of the B1 and B2 connectors where the inner conductor is standardized at 0·500 in diameter. The reasons for this choice are explained later. The outer conductors always form butt flange joints and are bolted together by three captive dowel screws. These screws are not shown fitted to the connectors, but a detail is given of the method of assembly in Fig. 1. The design allows initial finger tightening by means of the knurled head whilst final clamping is effected by a screwdriver or, if the instrument has a large front panel, by a spanner.

The inner conductors of a B2–B1 junction screw together directly. The immittance standards, which carry a B2 connector, embody a replica of a B1 connector at the remote ends so that a number of standards can be connected in series, as explained in Reference 1. The inner conductors of a B junction are joined by means of the single-ended plug, wh an A1–A1 or A1–A2 junction utilizes the double-ended In all cases the plane of cleavage is precisely defined by the c conductor flange ioint. When there is a change in characte impedance at a junction there will be a step in the inner ductor diameter at the plane of cleavage. The effect of step is to introduce a discontinuity capacitance which is culable and readily allowed for. A curve is referred Section 4.5 which enables the discontinuity capacitance i air-spaced line to be read off directly. It is important that lines are air-spaced for a minimum distance equal to the diffe between the coaxial conductor diameters, otherwise correc for the proximity effect of the dielectric supports would have made. The supports are not considered a part of the con system. This gives the instrument designer complete fre to minimize the discontinuity effect of the supports acceptable degree for the application in hand.

The coaxial connector line is air-spaced throughout its le the inner conductor being supported by the mating conn at each end. Immittance transformations through the lin therefore be made with high precision.

(3) USE OF THE SYSTEM

The system is intended primarily for use with standard precision laboratory instruments of all types in which immi is a major parameter, e.g. immittance standards, immi meters, wattmeters, insertion-loss meters, dissipative attenu directional couplers, resistive terminations, etc. The immi characteristics of all these devices have to be referred bac the immittance meter and standards to length and time mea ments. This means that a large number of intercompar have to be made, which emphasizes the fact that the u tainties introduced by the connector system must be compared with the ultimate precision needed in determ the immittance/frequency characteristics of the devices. example of this is given in Fig. 2, which shows the various s of admittance measurement necessary to design and calibr particular type of multi-range dissipative wattmeter. admittance bridge is first calibrated by means of the suscep standards, which are referred to length and time measurem This process has already been described.[1] The next stage i evolution of a satisfactory wide-band resistor-mount desig theory for which has to be verified by means of the admit bridge. Such a design, employing a tractorial resistor ja has been described by Harris.[2] This design is then used fo dissipative element of the milliwattmeter (basic range). voltage-measuring element of this instrument naturally a the admittance characteristic, so that a compensating net has to be employed. This means that the overall admit characteristic has to be verified by means of the admit bridge. The voltage calibration of the instrument is eff in terms of the standard cell in two stages in the manner sh The milliwattmeter (basic range) is made into a multi-r wattmeter by means of either dissipative attenuators or c tional couplers. The design of these instruments require theory of resistor-mount design and confirmatory admit measurements. In addition, the final calibrations are eff by means of the insertion-loss meter. This instrument requires similar design information and calibration faciliti order to ensure that the matched source and load, bet which the unknown is inserted, have acceptable admittanc quency characteristics. The piston attenuator embodied th is also standardized in terms of length measurements, i.e. diameter and axial displacement.



2.—Stages in the design and calibration of a wattmeter (multi-range).

(4) IMMITTANCE METER TERMINAL TYPE B1

(4.1) General Design Considerations

t any given frequency it is necessary, for practical reasons, nploy both an impedance and an admittance bridge to cover complete range of immittance encountered in practice. The er measures from substantially zero impedance up to some mediate value, and the latter from substantially zero ittance up to an intermediate value which will be the orocal of the other. The actual value which represents this sion between the ranges of measurement of the two types ridge is discussed in Section 4.2.

is logical that a standardized design of terminal should be l for both types of bridge because this will enable a common of immittance standards to be used, as well as a standardized gn of mating fitting for measurement purposes generally. precise measurement of immittance at radio frequencies only be undertaken in a completely screened system, the most common of which is the cylindrical coaxial form, and e is no doubt that bridge terminals should be of this type. se two principles having been accepted, the other design ors can now be discussed.

he accuracy required of the standards for use with precision ges is of the order of 0·1%. This means that the design he terminals must be such that, when the standards are con-ed for measurement, the calculated values are not changed nore than a few parts in 10^4.

n the case of the impedance bridge, which requires the minals to be short-circuited for initial balance, the plane of short-circuit must define the plane of measurement to within or two thousandths of an inch. In contrast, the admittance lge requires the terminals to be open-circuited for initial nce. In order that the plane of measurement shall be the e for both types of bridge, the fringe capacitance of the n-circuited terminals must be determinable to an equivalent order of precision, namely two or three thousandths of a picofarad.

Furthermore, when mating connectors are attached, having a value of characteristic impedance different from the bridge terminals, the discontinuity capacitance thus arising must also be calculable to the same order of precision as the fringe capacitance.

One further consideration is based on the desirability of adopting a design which will be suitable for the complete frequency range for which coaxial systems are normally used, i.e. up to about 4 Gc/s. This requirement places a maximum limit on the diameter of the outer conductor for a given value of characteristic impedance. The critical wavelength for the H_{11} mode is approximately equal to the mean circumference of the inner and outer conductors multiplied by the square root of the relative permittivity of the dielectric.[3] A suitable diameter for the outer conductor is 0·75 in, which makes the critical frequency higher than 4 Gc/s for the lowest impedance normally encountered.

The simplest form of terminal which satisfies all the above requirements is a 'sawn-off' air-dielectric line with the outer conductor terminated in a circular earth plane. This is illustrated in Fig. 1, type B1. The outer conductor carries three tapped holes and the inner conductor a single tapped hole for attaching the mating fittings. The general form of the terminal and the outer diameter having been decided, the question of the optimum value of the characteristic impedance has still to be settled.

(4.2) Characteristic Impedance

The approximate range of measurement of precision impedance and admittance bridges is shown in Fig. 3. The value of

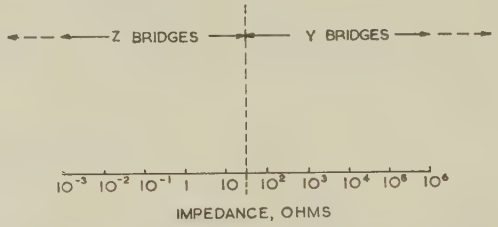


Fig. 3.—Range of measurement of impedance and admittance bridges.

impedance which forms the division between the two types of bridge is 30 ohms, i.e. the square root of one e.m.u. of inductance divided by one e.s.u. of capacitance.[4] This corresponds to a diameter ratio of $\sqrt{\varepsilon}$, i.e. 1·65.

When the precision dual admittance bridge[1] was being designed in 1947 it was apparent that the characteristic impedance of the coaxial standards, having the same dimensions as the bridge terminals, should not be higher than about 30 ohms, otherwise they would be excessively long for the required capacitance from both electrical and mechanical considerations. A low impedance favours good mechanical rigidity—a necessary condition because the standards were designed to be used horizontally and self-supporting on the terminals. Too low a value, however, would give rise to a mechanical problem of concentricity. Because the terminal impedance is not very critical, a choice was made in favour of a round figure for terminal diameter ratio rather than a round figure for impedance. The actual diameters adopted were 0·7500 in and 0·5000 in, giving a diameter ratio of 1·5000 and a limiting value of characteristic impedance of 24·30 ohms at the higher frequencies where the skin depth is negligibly small.

It is important that the dielectric is air for a minimum distance

of about 0·25 in from the end of the terminals in order that any higher modes excited by a discontinuity capacitance in the plane of measurement shall be sufficiently attenuated in the air-dielectric portion. If this were not the case, the discontinuity capacitances could not be calculated with sufficient accuracy.

(4.3) Fringe Capacitance

It has already been mentioned that, for an admittance bridge, the terminals are open-circuited at initial balance. This means that a correction must be applied for the fringe capacitance because this disappears when the unknown is attached for measurement. The approximate shape of the fringe field is shown in Fig. 4, from which it will be seen that it is mainly

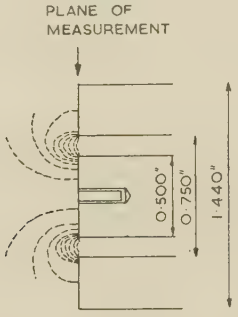


Fig. 4.—Terminal fringe capacitance field.

concentrated across the annular gap. The change in the fringe capacitance has been measured for different diameters of the earth plane. It was found that between diameters of 1·00–1·44 in the change was less than 0·002 pF, thus proving that a negligible number of lines of force extend to the edge of the earth plane. On the axis of the terminals the field also diminishes rapidly and is negligibly small at a distance of about 1 in from the terminal face.

The fringe capacitance has been determined by three methods, which will now be described. In a paper by Whinnery, Jamieson and Robbins[5] curves are given from which the discontinuity capacitance can be computed in a coaxial line with a discontinuous inner conductor [Figs. 7(*d*) and 11; also Figs. 7(*a*) and 8 in the Reference]. This condition can be reproduced at the B1 terminals by connecting to the outer conductor a cylinder of the same internal diameter. In practice the outer electrode of a standard coaxial capacitor is used. The change in capacitance caused by this connection will therefore be a measure of the transition from one state to the other. If the discontinuity capacitance for a discontinuous inner conductor is C_D and the measured change is ΔC, the terminal fringe capacitance $C_f = C_D - \Delta C$. The value of C_D obtained from the curves is $0{\cdot}407 \pm 0{\cdot}007$ pF, and the measured value of ΔC was $0{\cdot}066 \pm 0{\cdot}001$ pF, giving $C_f = 0{\cdot}341 \pm 0{\cdot}008$ pF.

The second method is by direct calculation employing a formula given in Reference 6. This formula, arranged in a form suitable for the case under consideration, is

$$C_f = \frac{0{\cdot}133\,7(a+b)}{\log_{10}^2 b/a}[\mathrm{E}(x) - 1] \text{ picofarads} \quad . \quad (1)$$

where a and b are in centimetres and $\mathrm{E}(x)$ is the complete elliptic integral of the second kind in which the argument is given by

$$(x) = \frac{2(b/a)^{1/2}}{1 + b/a} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

This formula gives a value of 0·345 pF for C_f.

In the third method the fringe capacitance is meas directly by means of a $\lambda/4$ short-circuited line. Such a presents zero susceptance to the terminals, and the differ between the bridge readings, with and without the line, is e to C_f. The experiment was carried out at 200 Mc/s using a constructed from silver tubes of better than 99·9% purity having the same dimensions as the terminals. The length o line was 37·436 ± 0·001 cm at 17° C and was based on a c puted velocity in the line of $2{\cdot}994\,85 \times 10^{10}$ cm/s at 17° C 30% r.h., the conditions obtaining at the time of the experim This velocity is appreciably less than the velocity[9] *in va* i.e. $2{\cdot}997\,93 \times 10^{10}$ cm/s, owing to the permittivity of air skin effect. The two correction factors are 0·999 70 0·999 27. The value of C_f obtained by this method $0{\cdot}344\,4 \pm 0{\cdot}003\,4$ pF. Precautions were taken to ensure the d.c. conductivity of silver was applicable at 200 Mc also measuring the conductance of the short-circuited line. value obtained was 51 ± 3 micromhos, compared wi theoretical conductance of 48·8 micromhos. Any signi difference in this respect would affect the accuracy of the ve correction factor due to skin effect. Allowance was also for the finite inductance and resistance of the brass sh circuiting unit, 0·38 pH and 0·48 milliohm respectively.

An estimation of the inaccuracy on C_f is difficult becau formulae from which the curves of Reference 5 were pre are not suitable for direct computation. The value of 0·4 for C_D was obtained by taking the mean of the two values by curves 8 and 11, there being a difference of 0·014 pF be the two results. It is probable that the second and third me are more accurate, but until further corroborative evi becomes available the following value has been adopted:

$$C_f = 0{\cdot}343 \text{ pF} \pm 1\% \quad . \quad . \quad . \quad .$$

At extremely high frequencies C_f will increase as the dimen of the terminals become comparable with the wavelength. for calculating the magnitude of this effect for the case u consideration are not given in Reference 5, but for the sp cases considered in Fig. 13(*a*) of the Reference it appears the change in C_f at 4 Gc/s is not likely to exceed about 5%.

(4.4) The Short-Circuiting Unit

The design of the short-circuiting unit is shown in Fi The short-circuiting element consists of a metal diaph approximately 0·01 in thick. This allows a small amou flexing should the two surfaces not be quite coplanar and ensures a sound electrical contact and prevents any u

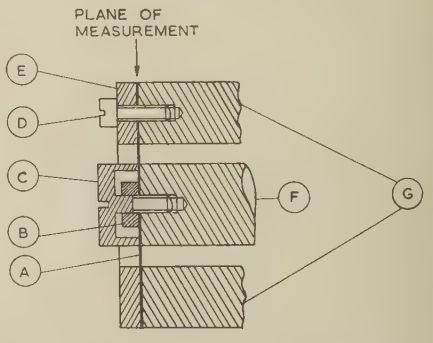


Fig. 5.—Details of short-circuiting unit.

A Diaphragm.
B Locating washer soldered to diaphragm.
C Cupped clamping screw.
D One of three screws.
E Clamping ring.
F Terminal inner.
G Terminal outer.

being placed on the terminals. Apart from being required e initial balance condition of an impedance bridge, it is sed as a short-circuit for the coaxial standards to provide ve susceptance standards for the admittance bridge (or e reactance standards for an impedance bridge).

short-circuiting annulus will have a finite impedance, but , sufficiently small to be ignored for most applications. ertain types of measurement it is, however, just significant. nula for calculating the impedance is given by Schelkunoff.[8] subscript f to denote the parameters of the annulus,

$$Z_f = R_f(1 + j) \quad . \quad . \quad . \quad . \quad . \quad (4)$$

fore

$$L_f = R_f/\omega \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

$$R_f = 0{\cdot}728 \times 10^{-4}(f\rho)^{1/2} \log_{10} b/a \text{ ohms} \quad . \quad . \quad (6)$$

ing in cycles per second.

ical values for R_f and L_f at 200 Mc/s, for a diameter ratio f 00, are given in the previous Section. It should be noted he reactance of the skin is always equal to the resistance, dition which holds in any system in which the skin depth ll.

(4.5) Discontinuity Capacitance

en measurements are made on a system having a charac- c impedance different from the bridge terminals, this will rise to a discontinuity which can be represented as a itance concentrated in the plane of cleavage. The capaci- can be derived from data given in Reference 5. It is d in Reference 1 for mating connectors of characteristic lances up to 120 ohms. At extremely high frequencies apacitance will increase as the terminal dimensions become arable with the wavelength. At a given frequency the ntage increase in C_d will be greater for the higher values . From Fig. 13(*a*) of Reference 5 it is estimated that at s the increase in C_d when $Z_c = 50$ ohms will be less than 1%. practice it is sometimes necessary to use a cone connector en the device being measured and the bridge terminals. ch instances the outer diameter of the cone at the bridge nals is made 0·750 in. Further information on cone cteristics and techniques is given in Reference 1. By oying the standardized design of connectors described is paper the use of cones is avoided in the design, ardization and calibration of precision laboratory instru- . When these instruments are subsequently used to cali- general-purpose measuring instruments of lower accuracy se of cones or adaptors is inevitable owing to the con- nal types of plug and socket employed on this class of ment. This situation will obtain until the designers of al-purpose instruments introduce a properly engineered ctor of standardized design.

(4.6) Terminal Resistance

e series resistance of the terminals is balanced out at initial ce with an impedance bridge and therefore does not give o errors when the unknown is measured in terms of a ce difference. For the admittance bridge, however, the nal resistance, if sufficiently large, will give rise to an in the conductance measurement when the unknown sus- nce is large. The form of the error term due to this cause $C_x^2 R_t$. The resistance of a coaxial conductor in which the lepth is small is given by Russell[10]:

$$R_t/l = \left(\frac{\rho f}{10^9}\right)^{1/2}\left(\frac{1}{a} + \frac{1}{b}\right) \text{ ohms/cm} \quad . \quad . \quad . \quad (7)$$

f is in cycles per second and a, b, and l are in centimetres.

The terminals of the dual admittance bridge described in Reference 1 were made sufficiently short for the resistance to be negligibly small for most conditions. This factor also had considerable bearing on the type B1 terminal design.

(4.7) Radiation Conductance

Because the terminals are initially open-circuited for the admittance bridge, an error can be caused in the conductance measurement if there is any significant radiation. A formula for calculating the radiation conductance of an open-ended coaxial line is given in Reference 7, which, rearranged in a more convenient form, is

$$G_f = 0{\cdot}252\left[f^2(b^2 - a^2)/\log_{10}\frac{b}{a}\right]^2 \text{ micromhos} \quad (8)$$

where f is in gigacycles per second and a and b are in centimetres.

The radiation conductance of the type B1 terminal is less than 1 micromho for frequencies below 830 Mc/s and is therefore negligible. At frequencies of several gigacycles per second the radiation conductance is significant and a correction would be necessary for any admittance measuring instrument operating in this range.

In this respect it is worth remembering that the standing-wave meter is really an immittance measuring instrument because it gives its optimum performance at that value which forms the division between the ranges which are most suited to impedance and admittance measurement (Fig. 3). Also, it has equal ranges of measurement into both the impedance and admittance regions. This means that its plane of measurement can be referred to either a short-circuit or an open-circuit. In practice it is more convenient to use the former method because it is more perfect in that the electrical and mechanical planes are coincident when the skin depth in the short-circuit is negligibly small. Apart from this it also eliminates any error due to radiation conductance.

It will be realized that the radiation conductance is the real portion of the admittance existing at the end of an open-circuited coaxial line and that the fringe susceptance forms the imaginary part of this admittance.

(4.8) Major Parameters

The major parameters of the terminals are the inductance and capacitance. Individually they give rise to the largest terminal corrections for admittance and impedance bridges respectively. For a diameter ratio of 1·500 the inductance is 811 pH/cm for an infinitely small skin depth and the capacitance is 1·373 pF/cm. The effect of finite current penetration at relatively low frequencies is to increase the inductance. The capacitance, however, is unchanged because the electric field does not penetrate the metallic surface to a significant degree. The fractional increase in inductance due to this cause can be obtained from a formula given by Russell[11]:

$$(1 + \zeta)^2 = 1 + \frac{\left(\frac{\rho \times 10^9}{4\pi^2 f}\right)^{1/2}\left(\frac{1}{a} + \frac{1}{b}\right)}{4{\cdot}605 \log_{10} b/a} \quad . \quad . \quad (9)$$

where f is in cycles per second and a and b are in centimetres.

To obtain the true inductance of the terminals the limiting value of 811 pH/cm must be multiplied by $(1 + \zeta)^2$. The correction is negligibly small at frequencies above about 200 Mc/s.

This correction is defined as the square of $(1 + \zeta)$ because $(1 + \zeta)$ is more frequently used as the correction on the limiting value of characteristic impedance and $1/(1 + \zeta)$ as the velocity correction factor due to skin effect for coaxial-line transformations. These corrections are considered in more detail in Section 8.

Although the major parameters of the terminals can be accurately calculated, this is not the case for the length of the terminals because the effective point of connection with the bridge-circuit elements is usually indefinite. This means that the total inductance and capacitance of the terminals must be measured. A method for doing this for an admittance bridge is described in Reference 13.

(5) IMMITANCE STANDARDS CONNECTOR TYPE B2

The method of assembly of a standard capacitor on the bridge terminals is shown in Fig. 6. It will be seen that the effective diameters of the standard are the same as the terminals; also, the remote end of the standard is a replica of the terminals. By

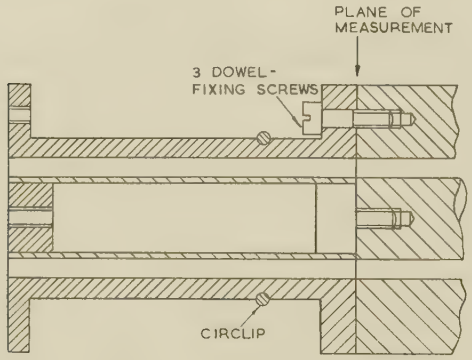


Fig. 6.—Standard capacitor attached to type B1 terminal.

this means two or more standards, having their lengths arranged in geometrical progression, can be joined in series to provide intermediate values. Because the fringe capacitance at the remote end is the same as that at the bridge terminals the change in capacitance is that due to a straight radial field and is therefore capable of precise calculation. The dimensions chosen are such that an appreciable degree of eccentricity can be tolerated without causing a significant change in capacitance. The capacitance of two cylinders of radii a and b and length l contained one within the other and with their axes parallel but displaced by a distance c is given by Russell[12]:

$$C = \frac{\epsilon l}{2 \cosh^{-1} A} \quad . \quad . \quad . \quad . \quad . \quad (10)$$

where $A = (b^2 + a^2 - c^2)/2ab$

and $\cosh^{-1} A = \log_\epsilon [A + \sqrt{(A^2 - 1)}]$.

The fractional change in capacitance due to this cause is shown in Fig. 7, from which it will be seen that an eccentricity of 0·003 1 in gives rise to a change of only 3 parts in 10^4. This degree of eccentricity can readily be achieved without resort to elaborate spigots and the like. With this design it has been found that the capacitance can be calculated to within 0·1% and that the constancy of the values of the standards is not worse than 0·03% over a period of years. The electrical characteristics of the standards are described in greater detail elsewhere.[1, 13]

(6) GENERAL-PURPOSE CONNECTOR TYPE A1

This connector is used for all types of instrument, other than immittance meters, in which immittance is a major parameter. The diameter of the inner conductor is made to suit the required value of characteristic impedance, usually 50 or 75 ohms. The former value is preferred because the mechanical design probl for extremely-wide-band instruments are much simpler ov to the larger diameter of the inner conductor. Furthermor it a relatively simple matter to change the impedance to 75 o by means of an impedance convertor comprising a series-i 25-ohm cylindrical film-resistor surrounded by a tractorial jac designed in accordance with Reference 2, and fitted with nectors of 75 and 50 ohms characteristic impedance at the e Such a convertor has a uniform frequency response from to several gigacycles per second. The conversion of a 75- basic system to 50 ohms is not so easy as it requires a s resistor of 150 ohms embodying a T-junction. The design 150-ohm coaxial resistor mount is far from optimum owin the large diameter ratio (12·3 : 1). This fact, coupled wit limitations imposed by the T-junction, restricts the frequ response to a much lower value than that which can be achi with the simple series arrangement.

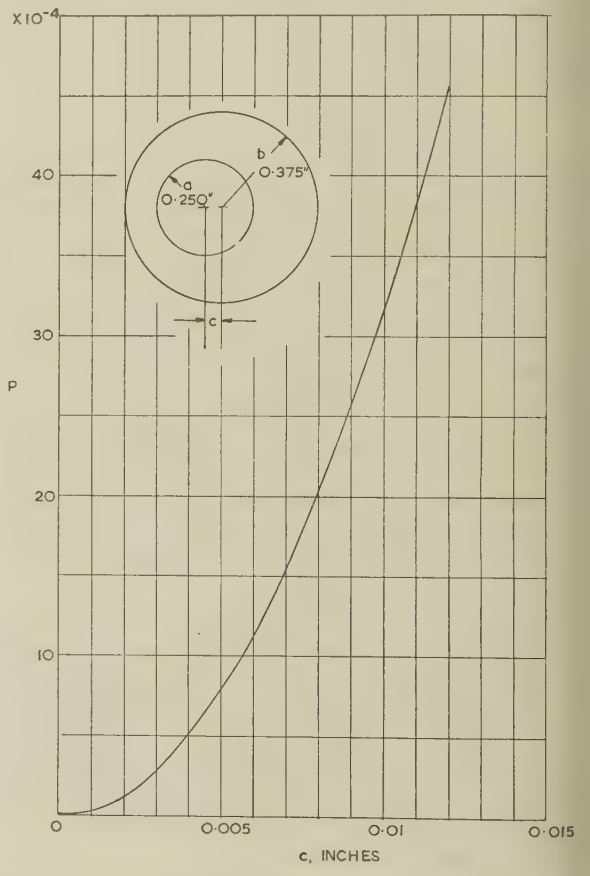


Fig. 7.—Fractional change in capacitance due to eccentricity betw electrodes of coaxial capacitance standard.

$$P = \left\{ \frac{\log_{10} [A + \sqrt{(A^2 - 1)}]}{\log_{10} b/a} - 1 \right\} \times 10^4$$

where $A = (a^2 + b^2 - c^2)/2ab$

As previously pointed out, the dielectric support for the i conductor must be situated from the flange face at a minim distance equal to the difference between the diameters of inner and outer conductors. Considerable care must be ta in the design of these supports in order that they are reflect less over the complete frequency range of the instrument. is best accomplished by undercutting the inner and outer ductors to a degree such that the inductance introduced fo with the two capacitances produced in the planes of the diele

ort surfaces, a π-section of substantially the same characic impedance as the line.[14, 15] The double undercutting ensures a satisfactory means of axially locating the inner ductor relative to the outer.

ue critical frequencies for the H_{11} mode for a 50-ohm conor are 7 and 5 Gc/s for dielectric permittivities of 1·0 and espectively. It should be remembered that it is the dielectric orts which first excite the higher unwanted mode.

(7) SIMPLIFIED CONNECTOR TYPE A2

certain types of instrument with input and output terminals, hich the power flow is in one direction only, the immittance acteristic at the output terminal is of no consequence and fore does not need to be measured. Such a device could n attenuator or the impedance convertor referred to in on 6.

will disappear and will be replaced by the discontinuity capacitance C_d. The admittance Y_i of the unknown, at the input to the connector line, in terms of the measured admittance Y_m will then be

$$G_i + j\omega C_i = G_m + j\omega(C_m + C_f - C_d) \quad . \quad (11)$$

For a short silver connector line the resistance can usually be ignored in the transformation of Y_i through the line to obtain Y_r. For example, at 200 Mc/s and for a 6 in connector line of 24·3 ohms characteristic impedance the errors arising through neglecting the resistance for an admittance of approximately $13·3 + j13·3$ millimhos is not greater than 0·04% on the conductance and 0·14% (0·015 pF) on the capacitance. In certain cases, however, the resistance may be significant, particularly at the higher frequencies. The resistance of silver lines of various characteristic impedances [eqn. (7)] is given in Fig. 9.

At frequencies below about 200 Mc/s the skin depth in the

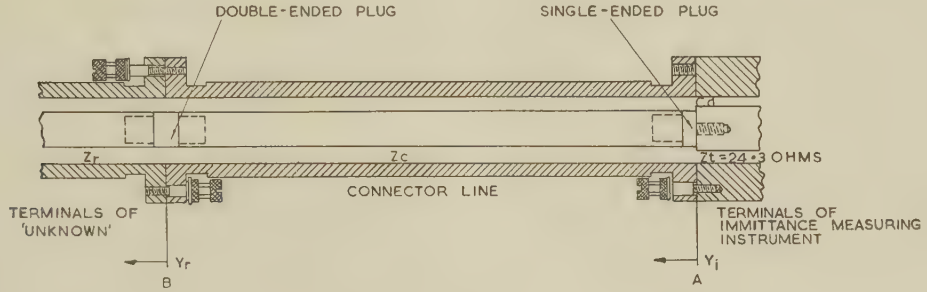


Fig. 8.—Practical measuring system using the connector line.

ne type A2 connector was introduced for this purpose. It is mplified version of A1 with the outer conductor formed h with the face of the instrument and, in consequence, is h shorter and less expensive. It connects directly to A1 will not mate with B1. However, it can be converted to by means of the connector line described in the next Section, uld this condition be needed for a special purpose.

(8) CONNECTOR LINE

he connector line is constructed from special silver-lined and r-overlay precision-drawn tubes. The silver coverings are uch purity and surface finish that the d.c. conductivity is licable in the calculation of skin depth up to the highest uencies. It is fitted with a type A1 connector at each end the inner conductor is supported by the mating connectors. special design feature of complete freedom from discon- ities is a necessary condition to allow precise admittance sformation through the line. The length of the line and the neter of the inner conductor are chosen to suit the require- nts. The only features which are standardized are the outer neter and the connectors.

is used where physical separation is necessary between two ruments and also to obtain accurate admittance trans- nations. In the latter application it is useful in standardizing lges when the line is used as a $\lambda/4$ transformer. A further is mentioned at the end of Section 7.

As an example of the use of the connector line consider it rposed between an unknown admittance, Y_r, and the type terminals of an admittance bridge (Fig. 8). At initial balance bridge terminals will be loaded with the fringe capacitance When the unknown admittance, with the connector line ached, is connected for measurement, the fringe capacitance

connector line is significant and must be taken into account when precise measurements are being undertaken. The effect of finite current penetration is to increase the inductance of the line because the effective diameters of the outer and inner conductors are increased and decreased respectively. This also

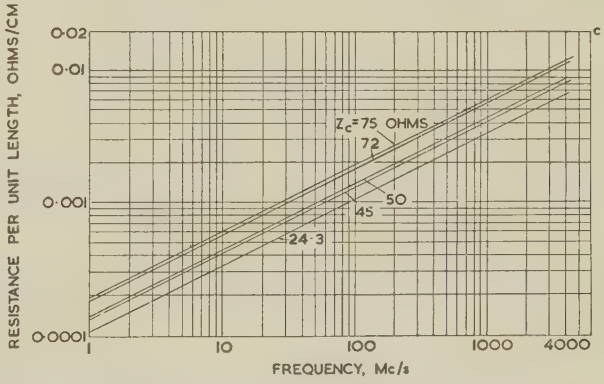


Fig. 9.—Resistances of silver coaxial connector lines of various characteristic impedances having an outer diameter of 0·750 in.

affects the velocity of propagation and the characteristic impedance but not the capacitance. If the inductance is increased by a factor $(1 + \zeta)^2$ [eqn. (9)], the characteristic impedance will be increased by the factor $(1 + \zeta)$ and the velocity reduced by $1/(1 + \zeta)$. If the limiting values of the parameters at very high frequencies, where the skin depth is substantially zero, are denoted by subscript ∞, then

$$L_\infty = 4\,605 \cdot 2l \log_{10} b/a \text{ picohenrys} \quad . \quad (12)$$

or $$L_\infty = l^2/v_\infty^2 C \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (13)$$

$$Z_\infty = 138{\cdot}02 \log_{10} b/a \text{ ohms} \quad . \quad . \quad . \quad (14)$$

or $$Z_\infty = l/v_\infty C \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (15)$$

where l = length of the line in centimetres

$$C = 0{\cdot}241\,76 l/\log_{10} b/a \text{ picofarads} \quad . \quad (16)$$

and $$v_\infty = 2{\cdot}997\,0 \times 10^{10} \text{ cm/s} \quad . \quad . \quad . \quad (17)$$

The values at lower frequencies where the skin depth is small but significant are therefore given by

$$v = v_\infty/(1 + \zeta) \quad . \quad . \quad . \quad . \quad (18)$$

$$L = L_\infty(1 + \zeta)^2 \quad . \quad . \quad . \quad . \quad (19)$$

$$Z = Z_\infty(1 + \zeta) \quad . \quad . \quad . \quad . \quad (20)$$

or $$Z = l/vC \quad . \quad . \quad . \quad . \quad . \quad . \quad (21)$$

The values of the constants in eqns. (14), (16) and (17) are based on an air dielectric at 20° C, 760 mm Hg and 50% r.h. They are not affected by more than one part in 10^4 for other values of humidity between zero and 100%. It should be noted that the line is defined by the capacitance, the velocity and the length. It has been found most convenient to calculate the capacitance using eqn. (16) and then to derive the characteristic impedance from eqn. (21) using eqns. (17) and (18). The value of ζ for silver lines having an outer diameter of 0·750 in and various characteristic impedances is given in Fig. 10. $(1 + \zeta)$

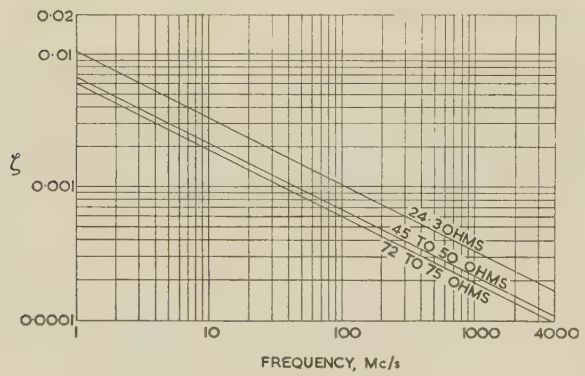


Fig. 10.—Reciprocal velocity factors $(1 + \zeta)$ for silver coaxial lines of various characteristic impedances having an outer diameter of 0·750 in.

is called the 'reciprocal velocity factor' and this is all that is needed to obtain the true velocities, inductances and characteristic impedances from the limiting values at substantially zero skin depth.

In order to calculate the unknown admittance, Y_r, in terms of the input admittance, Y_i, and the line parameters it is necessary to use the transmission-line equation because a circle diagram is not accurate enough for precise measurements. In order to simplify this process and avoid the use of complex functions the following formulae have been developed. These are quite simple to use because all the terms are dimensionless; they merely multiply the known conductance and effective line capacitance C' by factors to give the unknown conductance and capacitance.

$$G_r = \frac{G_i(1 + \tan^2 \beta l)}{\left(1 + \frac{C_i}{C'} \tan^2 \beta l\right)^2 + G_i^2 Z_c^2 \tan^2 \beta l} \quad . \quad . \quad . \quad (22)$$

$$C_r = \frac{C'\left[\left(\frac{C_i}{C'} - 1\right)\left(1 + \frac{C_i}{C'} \tan^2 \beta l\right) + G_i^2 Z_c^2\right]}{\left(1 + \frac{C_i}{C'} \tan^2 \beta l\right)^2 + G_i^2 Z_c^2 \tan^2 \beta l} \quad . \quad (23)$$

where v is given by eqn. (18), Z_c is given by eqn. (21), C $(C \tan \beta l)/\beta l$, where C is the total static capacitance of connector line as given by eqn. (16), and l = length of connector line in centimetres.

No approximations have been made in the derivation of th formulae from the transmission-line equation except the assu tion that the series resistance and shunt conductance of connector line are negligibly small.

In certain instances it may be necessary to express Y_i in te of Y_r. When this is so the following formulae are applicable

$$G_i = \frac{G_r(1 + \tan^2 \beta l)}{\left(1 - \frac{C_r}{C'} \tan^2 \beta l\right)^2 + G_r^2 Z_c^2 \tan^2 \beta l} \quad . \quad . \quad (24)$$

$$C_i = \frac{C'\left[\left(\frac{C_r}{C'} + 1\right)\left(1 - \frac{C_r}{C'} \tan^2 \beta l\right) - G_r^2 Z_c^2\right]}{\left(1 - \frac{C_r}{C'} \tan^2 \beta l\right)^2 + G_r^2 Z_c^2 \tan^2 \beta l} \quad . \quad (25)$$

The above two pairs of formulae can, of course, be use transform the admittance from one plane to any other pla a uniform coaxial system, but care must be taken to choos right pair. Eqns. (22) and (23) are applicable when tra forming in the same direction as the known admittance seen, and eqns (24) and (25) are used when transforming in opposite direction to which the known admittance Y_r is see

(9) CONCLUSIONS

A description has been given of a practical coaxial connec system for use up to 4 Gc/s with precision r.f. measuring inst ments. Such a system is essential when immittance is a ma circuit parameter. The errors arising in the measurement other quantities such as voltage, power, insertion loss, etc., to uncertainties in the immittance of a system based on design described would be insignificant, bearing in mind state of the art in the measurement of these quantities.

The system has been in frequent use at a Ministry of Aviat laboratory for many years and has proved satisfactory in ev way. It has also been adopted, or is being considered by number of national standards laboratories.

Reference has been made to the poor design of coaxial c nectors frequently used on general-purpose measuring inst ments. Although one or two manufacturers have apprecia the importance of this aspect and introduced special connect to minimize the discontinuity effect and eliminate the need male and female plugs and sockets, it is considered that situation is still far from satisfactory. The standardization connector design, employing the hermaphrodite principle, this class of instrument is still a pressing requirement. Its im mentation would constitute a very great step forward in standardization of r.f. measuring instruments. There is reason why such a connector system should not be made c patible with the precision connector so as to avoid the need cones, adaptors or any other interconnecting media. would not only lead to considerable improvement in accu in the calibration of general-purpose instruments aga precision laboratory instruments, but would also save time frustration by avoiding the need to design adaptors to meet requirements of each particular case.

The advantages of employing a low value of character

dance for coaxial measuring instruments have been pointed in Section 6. The obvious choice is 50 ohms. In this ct considerable advantage would accrue if this value was standardized.

roughout the paper reference is made to admittance and dance bridges. Although a suitable circuit and mechanical gn for a precision admittance bridge[1] have been found, the osite is the case for the impedance bridge. Impedance ge circuits either require elements in series or a variable tor in one of the arms. Either of these conditions rules out ossibility of achieving a precision bridge at several hundred acycles per second. No doubt the problem will be solved tually by employing a new technique or a new bridge circuit. he meantime the admittance bridge will satisfy all require- ts for precision measurements in relation to 50- and 75-ohm ial systems for voltage standing wave ratios down to 0·4 an uncertainty of 0·002 for frequencies up to several lred megacycles per second.

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THE PHASE VARIATION OF VERY-LOW-FREQUENCY WAVES PROPAGATED OVER LONG DISTANCES

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(*The paper was first received* 13*th June, and in revised form* 15*th October*, 1960.)

SUMMARY

The proposed use of very-low-frequency waves (10–20 kc/s) for navigational aids of the phase-comparison type has led to the study of the phase variations of these waves when propagated over distances up to 6000 km. The particular aspect of this study described in the paper is the relative phase variations over a pair of adjacent paths.

Measurements have been made of the variations in the phase difference between the signals received from a single transmitter at two spaced receiving stations. The two principal transmitters used were Varberg, Sweden, on a frequency of 17·2 kc/s and at a range of 1000 km, and Annapolis, United States, on 15·5 kc/s at a range of 6000 km. Receiver base-lines, up to 280 km in length, were situated in England and oriented both transverse to, and along, the direction of propagation. The observations were made at various seasons, and the results show the extent of the systematic diurnal changes and the random fluctuations: the former were of the order of 30° at the shorter range and 5° at the longer, and the latter had standard deviations at both ranges of the order of 4° in the day-time and 10° at night.

It is shown that the magnitude of the random phase fluctuations can be calculated from measurements of the corresponding amplitude fluctuations. In this way data on phase fluctuations at night at intermediate ranges and for receiver spacings up to 2000 km have been obtained. The relationship between the receiver separation and the correlation between fluctuations on two paths has been examined, and it is shown how the results of the phase measurements may be applied to other receiver arrangements.

(1) INTRODUCTION

During the past few years suggestions have been made for the use of very-low-frequency (v.l.f.) waves for long-range radio navigational aids. Examination of existing fundamental knowledge on the long-distance propagation of v.l.f. waves, particularly in regard to their phase variations, has shown the need for further investigations. Part of the work carried out at the Radio Research Station is described in the paper. Observations made by other investigators have been confined to fixed receiving positions:[1–3] the purpose of the Radio Research Station work has been to provide observational material at other ranges and of such a type that it can be more directly applied to the navigational problem of the comparison of signals propagated to a receiver from two or more separated transmitters. A hyperbolic type of navigational aid is envisaged using continuous-wave transmission, so that the chief characteristic of the waves to be investigated is the temporal and spatial variations of phase. The amplitude variations are mainly of importance in relation to the occurrence of deep minima at certain distances and times of day. A more detailed study of the amplitude variations can, however, provide data on the phase variations.

There are, generally speaking, two aspects of the phase variations which require investigation: the first is the absolute variations of phase of the wave over a single transmission path and the second the relative phase variations over a pair of paths. Studies of the first type have been carried out using a highly stable local oscillator as the phase reference at the receiving e of the path. This technique of measurement requires the tra mitter frequency to be of high stability (within about 1 p in 10^9) and the number of existing v.l.f. transmitters ha such a characteristic was limited, at the time of commencem of the investigations described in the paper, to a single stati GBR at Rugby. Thus, since it was desirable to make measurements in Great Britain and at long ranges from transmitters, most of the work was concentrated on the stu of relative phase variations in which transmitters of low stabil such as those at Varberg, Sweden, and Annapolis, Uni States, could be used.

The technique adopted for the investigation was to meas the variations in the phase difference between the signals recei from a single transmitter at two spaced receiving stations. T arrangement is the inverse of that of an actual navigati system, in which there would be two spaced phase-locked tra mitters and a single receiver, but the essential information be obtained more conveniently in this way. The measureme have been carried out over pairs of paths 100 to 6000 km length using receiver base-lines up to 300 km long and tra mitters operating in the frequency range 15–20 kc/s.

Amplitude recordings at a large number of sites at ranges to 7600 km from the transmitter have also been made. So of the results which are relevant to the phase investigation described in the paper, but the remainder will be publish separately. The application of both the phase and amplitu investigations to the assessment of the operation of a navigati aid will form the subject of a further paper.

(2) PHASE-MEASURING EQUIPMENT

The principle of operation of the equipment which is used measure the difference between the phases of the signals receiv at the two receiving sites is shown in Fig. 1. A full descripti of the equipment is to be found elsewhere.[4]

The phase difference between the v.l.f. signals is obtain indirectly by measuring the phase difference between two aud frequency signals, one of which is carried by telephone l from the remote station to the measuring station. The au frequencies are obtained by mixing the v.l.f. signals with refere signals derived from the pulse modulation of the v.h.f. tra missions received from the Gee navigational system. So long the phase relationship between the two reference signals rema constant, the measured phase difference will be the same as v.l.f. phase differences. The reference signals are derived fr the timing oscillators (75 kc/s) in the Gee receivers and are ph locked to the pulse repetition frequency of the Gee transmissio To ensure that phase shifts in the v.l.f. part of the equipment not cause errors, provision is made for injecting into the v. receiver at each station checking pulses having a harmonic the v.l.f. frequency to which the receivers are tuned. Th checking pulses are derived by frequency division from timing oscillators and are maintained in a fixed phase relati ship to the received Gee pulses.

Since the delay of an audio-frequency signal in the telepho

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The paper is an official communication from the Radio Research Station, Department of Scientific and Industrial Research.

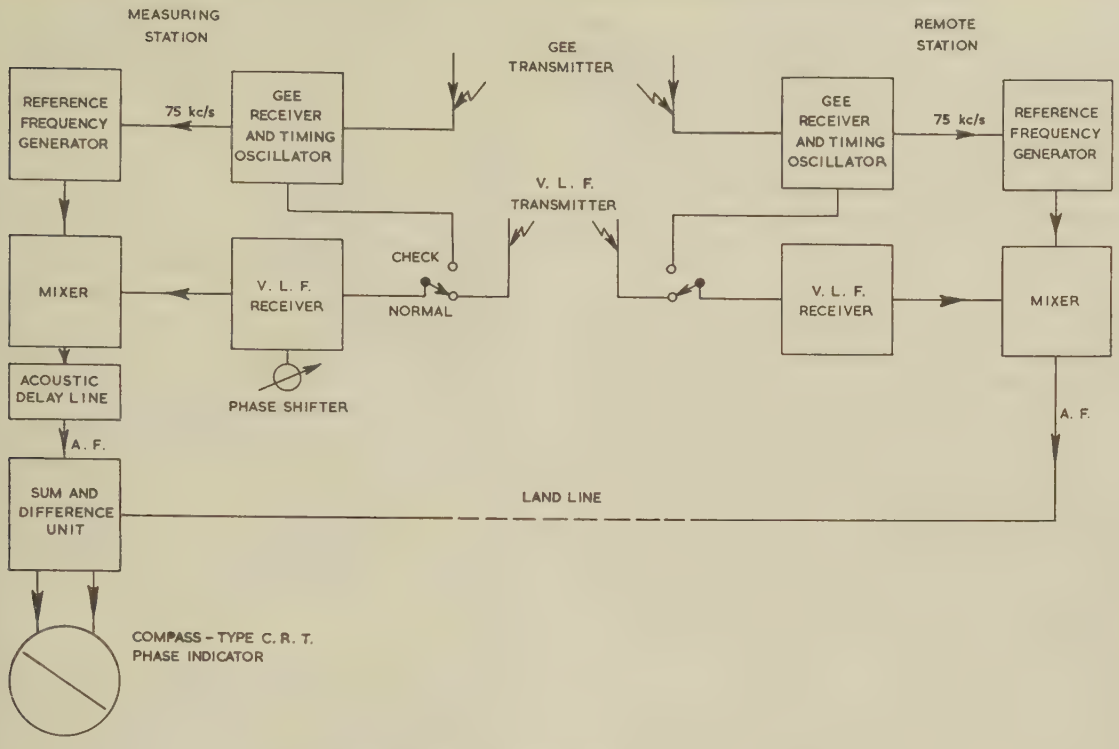


Fig. 1.—Basic system of measurement and layout of stations.

is appreciable, an acoustic delay line having a similar delay included in the equipment at the measuring station. This ures that keyed signals are applied simultaneously to the se-comparator unit and consequently allows the full duration he signals to be used for phase measurements. The inclusion he delay line also provides compensation for phase errors ing from small changes in the frequency of the transmitter.

The phase-comparator circuits consist of a sum-and-difference t followed by a compass-type cathode-ray tube. The inclination of the trace, relative to one of the axes, is equal to half the se difference between the signals.[5]

The major source of inaccuracies is poor signal/noise ratios at h the Gee and v.l.f. receivers. Measurements made in the oratory gave the holding accuracy of the Gee phase-locking cuits as equivalent to 1·2° of phase at 16 kc/s for a signal/noise io of 3 dB. With ratios of 12 dB or higher the accuracy is 1°. The effect of noise in the v.l.f. channel is to cause a ning out of the trace on the phase-indicator tube. The ding accuracy obtainable with a signal/noise ratio of 24 dB is)°, and for a ratio of 18 dB it is 2·0°.

The overall accuracy of the system under good signal/noise nditions is considered to be within ±2° at 16 kc/s for long-m measurements, while for periods up to 30 min the accuracy better than ±1° of phase.

(3) SHORT-DISTANCE MEASUREMENTS

(3.1) Description of Measurements

Three short series of phase measurements have been made on gnals from the Rugby transmitter, GBR (16·0 kc/s), with the easuring station at Slough and the remote stations at Slough, inkfield and Aldbourne successively (see Fig. 2). All three es were approximately 110 km from Rugby, and the spacings tween the receivers were 0·8, 10 and 75 km, respectively.

A vertical aerial and a loop set normal to the plane of propagation were employed at each station, and phase-difference measure-



Fig. 2.—Location of sites.

1. Slough.
2. Winkfield.
3. Teddington.
4. Cove.
5. Portsmouth.
6. Aldbourne.
7. Princes Risborough.
8. Banbury.
9. Cambourne.
10. Pembroke Dock.
11. Cambridge.
12. Hemsby.
13. Mablethorpe.
14. Carnaby.
15. Jodrell Bank.
16. Lower Whitley.
17. Greenock.
18. Inverness.
19. Swansea.

ments were made at 1 or 2 min intervals using the four possible combinations of aerials. That is, measurements were made of the relative phase between the signals received on vertical aerials at both stations, between those received on loop aerials and between those received on a vertical aerial at one station and a loop aerial at the other. By taking the differences between the values obtained the phase of the loop signal relative to that received on a vertical aerial at the same station could be deduced.

(3.2) Results

(3.2.1) General.

At a distance of 110 km from the transmitter the amplitude of the ground wave is about five times that of the one-hop ionospheric wave, and the ratio of the voltage induced in the vertical aerial by the ground wave to that induced by the ionospheric wave will be somewhat greater. The phase of the total output from this aerial can therefore be regarded as substantially that of the ground wave and used as a reference, while that from the loop aerial is dependent upon the phase of the received ionospheric wave. Thus changes in the phase difference between the loop and vertical-aerial signals at each station are a good approximation to the absolute phase changes of the ionospheric wave, while those between the two loops give its relative phase changes between the two receiving sites.

Phase changes of two main types are apparent from the measurements:

(*a*) Regular changes due to the diurnal changes of the ionosphere.
(*b*) Random variation having periods from a few minutes to an hour or more which are due to the irregularities of the ionosphere.

In the analysis of the results in this Section these types have been considered separately.

(3.2.2) Diurnal Changes.

Although only one record, that of 9–10th May, shows the complete phase changes from one day to the next, most of the others include the sunset period. The absolute and relative phase changes for this record, which was taken with the remote receiver at Aldbourne, are shown by the 10 min-mean plots in Fig. 3. It is seen that for this separation (75 km) the relative phase change at sunset is not discernible and that at sunrise it is small compared with the absolute phase change.

The observations at Winkfield and Slough also showed no marked sunset change of the relative phase, although the absolute phase change was of the order of 360°.

The points, ×, plotted in Fig. 3(*a*) are the hourly mean values of absolute phase recorded simultaneously at Cambridge by the Cavendish Laboratory. The detailed Cambridge record was compared with the corresponding Aldbourne one, and it was found that the relative phase variation between these locations, separated by 180 km and in mutually perpendicular directions from the transmitter, was of the same order as between Slough and Aldbourne [Fig. 3(*b*)]. It is of interest to mention that on another day (27th April), when a severe magnetic storm was in progress, the Cambridge record was markedly different from those taken at Slough and Aldbourne.

(3.2.3) Random Variations.

In order to examine the short-period variations of phase a statistical analysis of all readings (taken at 1 or 2 min intervals) was carried out. The day and night periods were treated separately, and for each period the standard deviation of the phase from the mean value was calculated. This was done for the absolute phase variations at each station as well as for the relative phase between the two. The results of this analysis are given in Table 1.

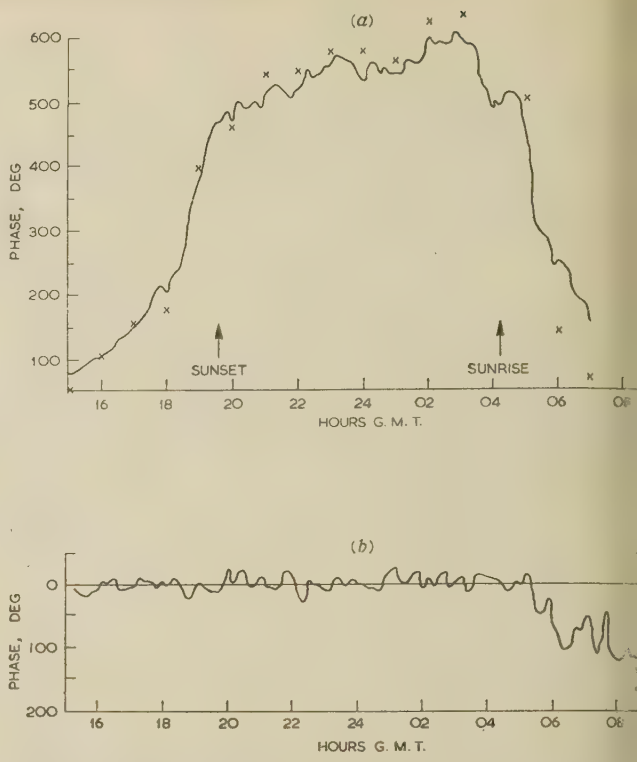


Fig. 3.—Phase variations over Aldbourne–Slough base-line.
Rugby transmitter, 9th–10th May, 1956.
(*a*) Absolute phase.
(*b*) Relative phase.
× Hourly mean values measured at Cambridge.

Table 1

STANDARD DEVIATION OF THE RANDOM PHASE FLUCTUATI OF THE HORIZONTALLY POLARIZED IONOSPHERIC COMPONE OF THE RUGBY SIGNAL

Series	Site separation	Total hours recording	Mean value of standard deviation			Perio
			Absolute phase		Relative phase	
			at Slough	at remote station		
	km		deg	deg	deg	
Slough–Slough	0·8	3·7	24·3	19·3	5·7	Nig
Winkfield–Slough	10	6·3	7·2 54·2	8·9 48·5	3·1 9·1	Da Nig
Aldbourne–Slough	75	8·3	5·5 24·0	3·5 25·5	2·7 13·0	Da Nig

Range 110 km

Although the amount of data used in the analysis is relati small one useful deduction can be made. There is a mar increase in the fluctuations during the night-time periods, average night value of the standard deviation of the abso phase being about 30° compared with a day-time average of 6

(4) MEDIUM-DISTANCE MEASUREMENTS

(4.1) General

Several series of measurements have been made of both phase and amplitude of signals received in this country from

mitter SAQ at Varberg, Sweden. The transmissions were odulated continuous wave on a frequency of 17·2 kc/s at a nal output power of 50 kW.

(4.2) Description of Measurements

ur series of measurements have been made with the uring station at Slough and the remote stations successively dbourne, Hemsby (Norfolk) and Lower Whitley (Cheshire) Fig. 2), two series being made at the latter site. The site emsby is on the line between Varberg and Slough at a ace of 203 km from the latter, while that at Lower Whitley proximately at the same range as Slough but situated m to the north. These last two sites were selected in this in an attempt to separate the random variations from the al changes. It seemed likely that the relative diurnal ges at these comparatively short receiver separations would preciably less when the receivers were at equal ranges than they were at different ranges on the same radial. The erse might also be true for the random variations.

these measurements either a vertical aerial or a loop set e plane of propagation was used for reception.

(4.3) Results

) General.

a range of about 1 000 km the ground and ionospheric s are of the same order of magnitude, and since the angle rival of the ionospheric wave is close to that of the ground , it is not practicable to separate out the latter and so obtain asurement of the absolute phase variations of the ionospheric as was done at the shorter ranges. The phase variations sured were therefore those of the total vertical component e field at one receiving station relative to those at the other. ough this may not be so satisfactory for the study of the spheric irregularities it closely resembles the conditions of ation of a navigational aid.

ie observations have again been analysed to show the nal changes and the random fluctuations separately.

(4.2) Diurnal Changes and Long-Period Variations.

ie set of curves in Fig. 4 shows the 40 min mean of the ive phase between the three pairs of stations. The dotted ions of the curves indicate periods of unreliable or missing rvations.

ie Aldbourne–Slough measurements, which are shown in the four curves, were made during the period 10th April–3 May, 1956. There are very marked differences between e curves, particularly at sunset, where there is a progressive ge within 8 days from an increase of over 60° of phase to crease of 40°. The very much steadier conditions in the time than at night are also apparent.

he Hemsby–Slough series, which were made in August, 1956, v a more definite diurnal change with a greater, but more able, phase difference at night than in the day-time. The marked changes that occurred between 0100 and 0500 hours the 9th August are to be noted and also the tendency for n to be repeated, but with reduced amplitude, on the 1 August.

he last five curves relate to measurements taken over the ver Whitley–Slough base-line during November, 1956, and , 1957. In these the diurnal change is small but the night- variations are considerable. An ionospheric storm was in gress during the last two periods (12–15th November), which unts for the large variations throughout both records. The curves show very steady day-time conditions, although e phase changes occur during the night. The diurnal nges are again small.

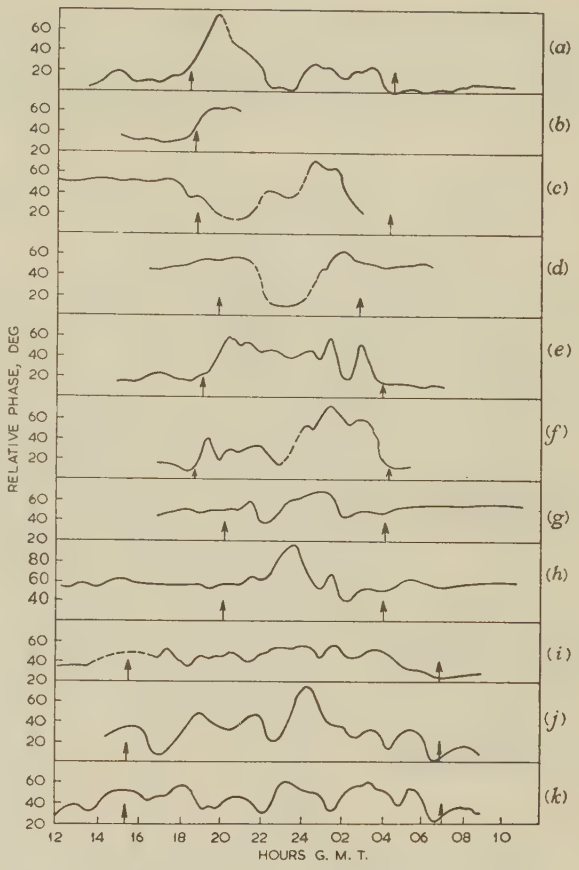


Fig. 4.—Mean relative phase: Varberg transmitter.

Aldbourne–Slough: (*a*) 11th–12th Apr., 1956. (*b*) 16th Apr., 1956. (*c*) 18th–19th Apr., 1956. (*d*) 30th–31st May, 1956.

Hemsby–Slough: (*e*) 8th–9th Aug., 1956. (*f*) 22nd–23rd Aug., 1956.

Lower Whitley–Slough: (*g*) 18th–19th July, 1957. (*h*) 22nd–23rd July, 1957. (*i*) 9th–10th Nov., 1956. (*j*) 12th–13th Nov., 1956. (*k*) 14th–15th Nov., 1956.

Arrows indicate mid-path sunrise and sunset times.

In order to obtain some quantitative comparison between the results over the three base-lines an analysis of the data was made in the manner shown in Table 2.

The Aldbourne–Slough results show large variations from day to day and are of little value for comparison purposes. The results over the two other base-lines, however, show that the general change of relative phase from day to night is less when the receivers are at equal distances from the transmitter than when they are on the same radial. The exceptionally large change of 21·9° recorded for 22nd–23rd July in the Lower Whitley–Slough series is explained by reference to Fig. 4. It will be seen that over the night period analysed (2300–0140 hours) there is a large increase in the relative phase which results in an abnormally high mean night value. All phase plots in this series showed large changes during the night period, but on this particular night the changes were exceptionally large and were positive over the whole analysis period.

Comparison between the November and July measurements over the Lower Whitley–Slough base-line shows that apart from the greater mean level of the relative phase during the summer month there is no significant difference due to season. The maximum deviations of the long-period variations from their

Table 2

Diurnal Changes of Relative Phase of Varberg Signals

Series	Date	Record length	Overall change of phase	Relative phase				Night mean minus day mean
				Day		Night		
				Mean	Maximum deviations	Mean	Maximum deviations	
		hours	deg	deg	deg	deg	deg	deg
Aldbourne–Slough April, 1956	11	14	80	86·5	−1 +3			
	12			83·5	−9 +5	99·0	−19 +24	+15·5
	16	6		52·4	−2 +6			
	17			38·9	−3 +3			
	17			38·1	−4 +3			
	18	15	60	61·9	−1 +2	(56·4	−20 +31)	(−5·5
May, 1956	30	14	(55)	122·5	−4 +2	(88·3	−8 +12)	(−34
Hemsby–Slough August, 1956	8–9	16	40	55·9	−3 +4	86·6	−9 +14	30
	9	8		53·4	−2 +3	(80)		(27)
	21	4½	—	54·9	−5 +6			
	22	13	70	54·0	−5 +3	69·0	−16 +14	
	22–23					84·4	−15 +15	30
						99·9	−15 +16	
Lower Whitley–Slough November, 1956	7–8	5	—			50·1	−11 +12	
	9–10	17	35	35·5	−6 +7	47·7	−17 +13	12·2
	12–13	19	72			67·8	−26 +43	
	14–15	23	50	40·9	−7 +7	55·4	−17 +18	14·5
Lower Whitley–Slough July, 1957	16–17	16	60	62·5	−3 +3	72·4	−27 +27	9·9
	18–19	18	37	51·5	−2 +4	61·1	−16 +17	9·6
	21–22	14	25	66·0	−5 +4	68·0	−10 +12	2·0
	22–23	24	65	58·6	−4 +5	80·5	−27 +20	21·9
				58·0	−6 +5			

Range about 1 000 km.
Figures in brackets are open to doubt as they are taken from records which were of short duration or discontinuous.

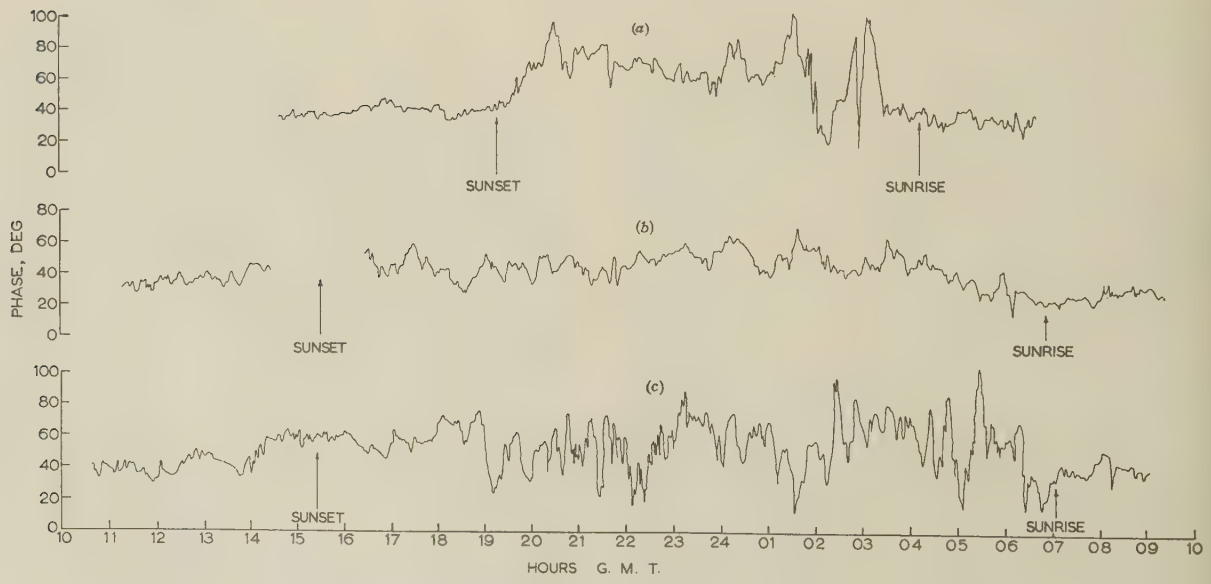


Fig. 5.—Relative phase variations: Varberg transmitter.

(*a*) Hemsby–Slough, 8th–19th Aug., 1956.
(*b*) Lower Whitley–Slough, 9th–10th Nov., 1956. Quiet ionospheric conditions
(*c*) Lower Whitley–Slough, 14th–15th Nov., 1956. Disturbed ionospheric conditions

day and night values emphasize the greater amplitude of variations at night than in the day, but show no significant ion with orientation of receiver base-line or with season.

he fourth column of Table 2 is given the overall change of e phase occurring during the total period of the record this covers a sufficient portion of the 24 hours for the s to be significant. These show larger variations from day y than from one base-line to another or from one season other. Over the longer base-lines the average value is 50° variations from 25° to 70°.

Random Fluctuations.

mplete curves drawn through all the measured points for typical days are shown in Fig. 5. Curve (*a*) was taken on Iemsby–Slough pair of stations and shows the marked se in amplitude of the rapid fluctuations that occurs during ight-time. Curves (*b*) and (*c*) show the results obtained e Lower Whitley–Slough pair for two periods separated ur days. Curve (*b*) shows the day and night fluctuations normal day while curve (*c*), which was taken during dis- d ionospheric conditions, shows greatly increased fluctua- at night. The day-time variations, however, were of the order on both days.

analysis of all the observations obtained on the Varberg mitter is given in Table 3.

e data are divided into day and night periods, having ions as shown, and excluding the sunrise and sunset periods. agnitudes of the fluctuations are expressed in terms of the tions from the mean for each period as given in Table 2. computations were based on phase readings at 2 min vals. It will be noted that although there are substantial ges from day to day there is a definite increase in the nitudes of the fluctuations at night, compared with the day- The standard deviations are also greater over the larger base-lines between Hemsby and Slough (200 km) and between Lower Whiteley and Slough (250 km) than over the shorter Aldbourne–Slough base-line (75 km). There seems to be no significant change of magnitude with season or with base-line orientation.

(5) LONG-DISTANCE MEASUREMENTS

(5.1) Description of Measurements

A few relative phase measurements were made on transmissions from Annapolis (19·0 kc/s) at a range of 5 800 km using the Aldbourne–Slough pair of measuring stations and the Hemsby–Slough pair.

One complete record for the night period between ground sunset at Slough and ground sunrise at Annapolis was made for each of the pairs of stations; on the Aldbourne–Slough pair in June, 1956, and on the Hemsby–Slough pair in August, 1956.

The main series of measurements on Annapolis transmission, however, was made with the measuring station at Slough and the remote station at Carnaby, Yorkshire, over a base-line of 286 km oriented almost normal to the direction of propagation (see Fig. 2). The observations were made over six periods of between 14 and 22 hours' duration in the months of February and March, 1958. In 1957 the operating frequency of the Annapolis station was changed from 19·0 kc/s to 15·5 kc/s so that this main series of measurements was made on the lower frequency.

(5.2) Results

(5.2.1) General.

The results of the long-distance measurements were analysed in the same manner as were those for the medium distances.

Table 3

RANDOM RELATIVE PHASE FLUCTUATIONS OF VARBERG SIGNALS

Series	Date	Day		Night		Average standard deviation	
		Standard deviation	Period	Standard deviation	Period	Day	Night
		deg	hours	deg	hours	deg	deg
Aldbourne–Slough April, 1956	11–12	4·7	5½	6·0	3	2·9	5·2
	16	2·8	3				
	17	2·5	5				
	18–19	1·2	4½				
May, 1956	30–31	3·1	2	4·5	1½		
Hemsby–Slough August, 1956	8–9	4·6	4½	7·9	4	3·6	10·7
		3·8	2				
	9	2·4	3½				
	21	3·9	3½				
	22–23	3·1	2	13·5	7½		
Lower Whitley–Slough November, 1956	7–8			8·7	5	4·1	12·4
	9–10	4·5	3	6·9	11½		
		2·3	1½				
	12–13			18·4	11½		
	14–15	5·0	3½	15·5	12		
		4·7	1				
Lower Whitley–Slough July, 1957	16–17	3·0	6	15·0	2½	3·7	11·8
	18–19	3·2	6	12·2	2½		
	21–22	3·8	4	6·0	2½		
	22–23	4·7	7	14·0	2½		
		3·8	6				

Range about 1 000 km.

(5.2.2) **Diurnal Changes.**

The diurnal and long-period changes in relative phase are illustrated by the curves in Fig. 6 and the analysis in Table 4.

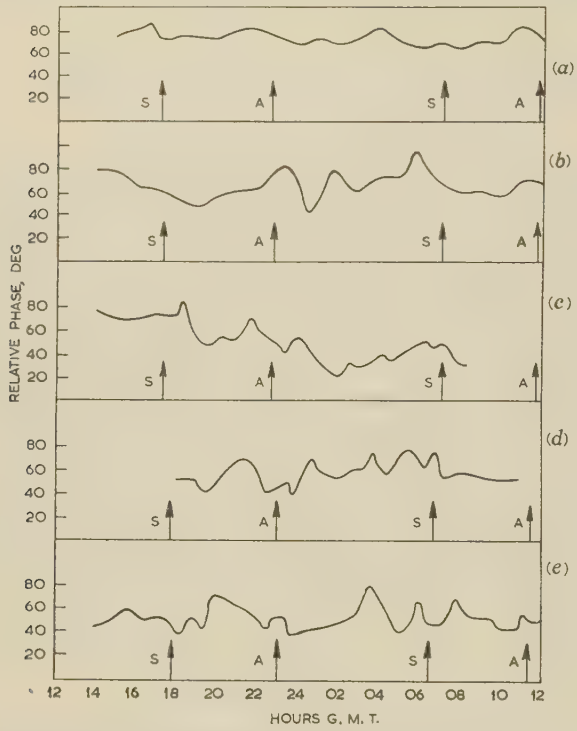


Fig. 6.—Mean relative phase: Annapolis transmitter.

Carnaby–Slough.
(*a*) 15th–16th Feb., 1958.
(*b*) 17th–18th Feb., 1958.
(*c*) 19th–20th Feb., 1958.
(*d*) 4th–5th Mar., 1958.
(*e*) 6th–7th Mar., 1958.
S. Sunrise and sunset times at Slough.
A. Sunrise and sunset times at Annapolis.

Since the sunset times at Annapolis and Slough diffe about 5 hours there were appreciable periods when the prop tion path was partly in daylight and partly in darkness. T fore, for the purpose of this analysis, the 24-hour period been divided into four:

(*a*) Day—whole path in daylight.
(*b*) Sunset—sunset line across path.
(*c*) Night—whole path in darkness.
(*d*) Sunrise—sunrise line across path.

Apart from showing that the overall deviation from period mean value is dependent upon the orientation and le of the base-line, the Aldbourne–Slough and Hemsby–Sl observations contribute little to the information on p changes at this range. The main facts to be obtained fro analysis of the Carnaby–Slough series (see last three lin Table 4) are that the change of the mean phase from da night is on the average small and in a downward direc although the changes between the sunset period and the da night periods may be greater. The maximum deviations f the mean of each period are again greater at night than i day-time, and the overall phase change throughout the da a mean value of 34° with variations from 20° to 47°.

(5.2.3) **Random Fluctuations.**

A complete curve drawn through all the measured poin a typical day is shown in Fig. 7, and the standard deviatio the fluctuations from the mean value for each period of the are given in Table 5.

The analysis again shows an increase in the magnitude o fluctuations during the hours of darkness in all cases. series of measurements over the Carnaby–Slough base-line sh a wide variation from day to day which seems to be ra greater in the day-time than at night. In view of the variat shown over this base-line no reliable deductions can be m from the one day's observations on the two other base-line to the dependence of the magnitude of the fluctuations on base-line length or orientation. From the medium-dist

Table 4

DIURNAL AND LONG-PERIOD CHANGES OF RELATIVE PHASE OF ANNAPOLIS SIGNALS

Series	Date	Record length	Overall change	Relative phase							
				Day		Sunset		Night		Sunrise	
				Mean	Maximum deviation	Mean	Maximum deviation	Mean	Maximum deviation	Mean	Maxim deviati
		hours		deg	deg	deg	deg	deg	deg	deg	deg
Aldbourne–Slough June, 1956	13–14	13	30	—	—	65·5	−8 + 7	76·9	−7 + 9	65·5	− 8 +
Hemsby–Slough August, 1956	20–21	15	55	—	—	23·7	−19 + 16	37·2	−17 + 25	26·0	−12 -
Carnaby–Slough February, 1958	15–16	21	23	80·0	−7 + 8	77·0	−4 + 5	73·0	−6 + 9	72·6	−7 +
	17–18	22	45	71·5	−8 + 8	58·0	−8 + 7	71·7	−23 + 18	61·0	−4 +
	19–20	14	47	71·8	−4 + 4	57·5	−8 + 10	62·5	−17 + 15	—	—
March, 1958	2–3	17	20	65·7	−1 + 1	57·8	−8 + 11	60·0	−11 + 6	—	—
	4–5	17	35	—	—	54·7	−9 + 10	61·1	−17 + 15	54·9	−4 +
	6–7	22	35	50·9	−6 + 7	55·2	−15 + 16	50·6	−12 + 20	52·0	−7 +
Averages ..			34	67·5	11	60·3	18	63·2	28	60·1	15
Average changes of means					−7		+3		−3		
Average day-to-night change						−4					

Range about 6 000 km.

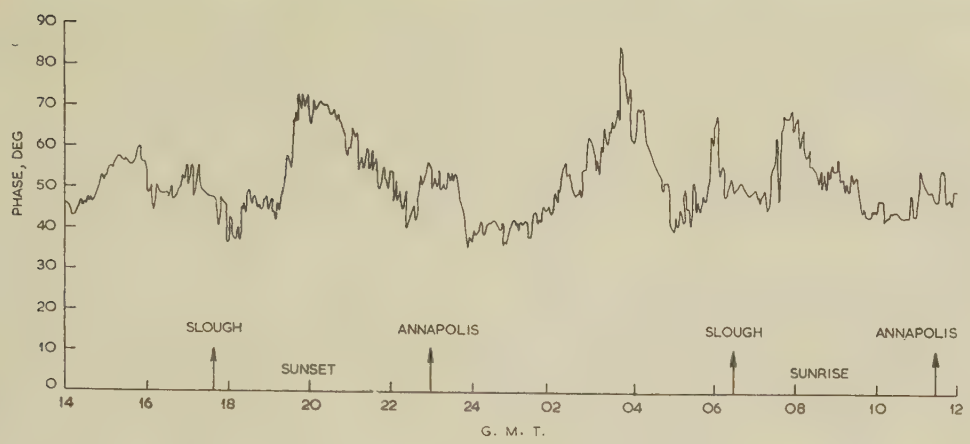


Fig. 7.—Relative phase variations: Annapolis transmitter.
Carnaby–Slough, 6th–7th Mar., 1958.

Table 5

RANDOM RELATIVE PHASE FLUCTUATIONS OF ANNAPOLIS SIGNALS

Series	Date	Day		Sunset		Night		Sunrise		Transmitter frequency
		Standard deviation	Period	Standard deviation	Period	Standard deviation	Period	Standard deviation	Period	
		deg	hours	deg	hours	deg	hours	deg	hours	kc/s
bourne–Slough une, 1956	13–14	—	—	5·0	4	7·2	2	5·8	5	19·0
nsby–Slough ugust, 1956	20–21	—	—	11·3	5	16·8	6	7·6	4	
naby–Slough ebruary, 1958	15–16	4·3	3	2·6	4	5·1	8	5·3	5	
	17–18	6·6	3	5·8	5	10·6	8	6·4	5	
	19–20	2·5	3	6·5	4	10·1	8	—	—	15·5
March, 1958	2–3	1·2	3	6·5	4	5·7	8	—	—	
	4–5	—	—	7·5	4	10·0	8	4·5	4	
	6–7	4·3	$3\frac{1}{2}$	10·8	$4\frac{1}{2}$	9·8	8	6·8	$4\frac{1}{2}$	
	Average	3·8		6·6		8·5		5·8		

Range about 6 000 km.

lts and other considerations it would be expected that the nitudes would decrease with decrease of base-line length.

(6) DISCUSSION OF RESULTS

eviewing generally the results of the phase measurements we that the data obtained over the short paths (100 km), ough of limited application to the navigational problems, new information on the relative phase fluctuations on cent ionospheric paths.

he measurements over paths of the order of 1 000 km give e indication of the dependence of the relative phase varia- s on base-line orientation, base-line length and season. The ition of observations over the longer paths of about 6 000 km ble the effect of path length on the relative phase variations e studied over base-lines of comparable length and approxi- ely the same orientation with respect to the transmitters. measurements give data on the change in the magnitude of random phase fluctuations from day to night.

lowever, as pointed out in Section 1, a comprehensive ly of the phase variations requires the observations to be iciently extensive to embrace a wide range of transmitter- and receiving-station arrangements and of ionospheric conditions, and it is clear that the work described above has done no more than obtain samples. As was realized at an early stage, the method of measurement has been responsible mainly for this limitation because of the considerable experimental effort required to obtain the observations and the restriction on the base-line length imposed by the effective range of the v.h.f. reference signals. It was decided, therefore, to supplement the phase observations by amplitude measurements both at the sites used for the phase-measuring equipment and at a number of other sites. Equipment for amplitude measurements is very much less complex than that for phase measurements, can be readily duplicated and can be made to operate practically unattended. A ready means is thus provided to enlarge both the temporal and spatial extent of the observations, although it must be realized that the data provided by the amplitude measurements are of limited application to the main navigational-aid problem.

The amplitude measurements are described in the following Sections and the results which are relevant to the study of relative phase variations are given.

(7) MEASUREMENT OF SIGNAL AMPLITUDE

(7.1) Introduction

The value of amplitude measurements in the study of v.l.f. propagation in relation to navigational aids is threefold:

(*a*) The determination of the variation of the general field-strength level with distance and time.
(*b*) Their use in conjunction with the phase measurements for deducing the characteristics of the ionosphere.
(*c*) The derivation of random phase fluctuations from the random amplitude fluctuations.

The first is of importance for predicting the existence or otherwise of regions of low field strength which might lead to unreliable operation of the aid due to low signal/noise ratio or, what is probably more important, the introduction of a false phase indication. This latter fault may arise when the low field strength is due to interference between two or more waves and the resultant signal undergoes a phase change of one or more complete cycles.

In the paper only the results of the analysis of the random fluctuations of amplitude are discussed since this aspect of the amplitude study is the one more closely related to the relative phase measurements described above. The diurnal amplitude changes and their relation to the absolute phase changes as well as the variation of field strength with distance will form the subject of a further paper.

(7.2) Description of Measurements

The amplitude-measuring equipment was of conventional design and produced a continuous pen record of amplitude. Superhetrodyne receivers with bandwidths of 50–150 c/s were used and their gain stability was such that a daily calibration was adequate to maintain an overall accuracy of measurement of better than ± 1 dB. Short vertical aerials of about 6 m in height were used, and the overall calibration of the equipment in terms of the strength of the vertical component of the field was made at all sites using a portable field-strength measuring set.

Recordings were made of the signals from the transmitters at Rugby, Criggion, Varberg and Annapolis. A large number of receiving sites in England, Wales and Scotland were used, and receiving equipment was also installed at Malta and in two of the Ocean Weather ships at the Meteorological Office. These two ships were stationed in rotation in four localities in the North Atlantic and observations were made during the passages to and from these localities as well as while they were on station.

The recordings were started in August, 1956, and, although not continuous, covered most seasons of the year until March, 1958. The observations were made at several sites simultaneously, the number varying from two up to six. The disposition of the sites is shown in Figs. 2 and 8, and details of the transmitters, ranges and periods of observation are shown in Table 6.

It will be seen that the frequency range covered was from 15·5 to 19·6 kc/s, the path lengths were from 100–5 900 km and the separations between receiving sites at which simultaneous observations were made were from 50 to 2 200 km.

(7.3) Analysis of the Random Fluctuations

In the analysis of the random fluctuations of amplitude and their relation to the phase fluctuations, the concept of a random signal described by Ratcliffe[6] and others[7,8] is used. It is assumed that the total signal is represented by a steady component, plus a random one, whose amplitude has a Rayleigh distribution and whose phase angle relative to that of the steady component has all values with equal probability. Provided that the random component is much smaller than the steady one, the standard

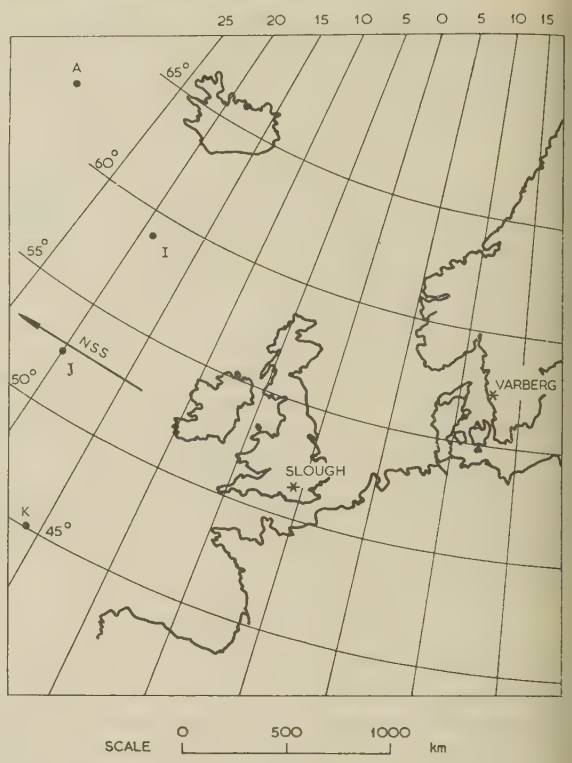


Fig. 8.—Location of weather-ship stations in North Atlanti

deviation of phase in radians is equal to the standard devia of the amplitude expressed as a fraction of the mean ampli It has been shown that the spatial correlograms of the phase amplitude are identical.

The practical importance of these results is that the abso phase fluctuations over a path or the relative phase fluctuati over a pair of paths can be readily calculated from amplit measurements over these paths. In addition, the correla between the phase fluctuations over adjacent paths can calculated for various path separations, and hence the le of base-line between receiving sites at which the correlation to a negligible value can be determined.

However, in applying this concept of a signal composed steady and a random component to v.l.f. signals propag over long distances, one must recognize the fact that the so-ca steady component will show systematic variations owing to diurnal changes in the ionosphere. It is necessary, therefore confine the analysis to periods which are clear of these system changes, and, in practice, only the night-time periods clea the sunrise and sunset effects were used.

In order to check the basic assumptions concerning the na of the signals, the standard deviations of the relative phas variations between a pair of receiving stations were compu from the corresponding pair of simultaneous amplitude rec ings, and the results were compared with the measured valu relative phase. If the observations at the two stations, 1 an separated by a distance d, give amplitude fluctuations of ma tudes a_1 and a_2 and are related by a spatial correlation coeffic $\rho_a(d)$, the standard deviations of the corresponding p quantities can be computed from the relations

$$\sigma_{\phi 1} = \sigma_{a1} \qquad \sigma_{\phi 2} = \sigma_{a2} \qquad \rho_\phi(d) = \rho_a(d)$$

and the standard deviations of the difference of phase from

$$\sigma^2_{\phi 1-2} = \sigma^2_{\phi 1} + \sigma^2_{\phi 2} - 2\sigma_{\phi 1}\sigma_{\phi 2}\rho_\phi(d)$$

Table 6

SITES AND RANGES AT WHICH AMPLITUDE AND PHASE MEASUREMENT WERE MADE

Transmitter	Observation period	Amplitude			Relative phase		
		Sites	Ranges	Separations	Sites	Ranges	Separations
			km	km		km	km
by 16·0 kc/s ..	Mar.–May, 1956				1, 2	110	0·8, 10
					1, 6	110	75
berg 17·2 kc/s ..	Apr.–May, 1956				1, 6	1044	75
	Aug., 1956	1, 12, 5	841–1121	83–284	1, 12	841, 1044	206
		1, 5, 11, 12	841–1121	83–284			
	Nov., 1956	1, 16	1038, 1044	245	1, 16	1038, 1044	245
		1, 7, 8, 16	1038–1046	32–245			
	Jan.–Feb., 1957	1, 12, 15	841–1044	206–275			
		1, 12, 15, 11, 3, 9	841–1392	16–395			
	Apr., 1957	1, 10, 12	841–1269	206–468			
		1, 10, 12, K	841–2371	206–1520			
	July, 1957	1, 16, 17	1038–1062	245–555	1, 16	1038, 1044	245
		1, 12, 16 17, J	841–2112	206–1420			
	Oct., 1957	1, 17	1044, 1062	555			
		A, K	2371, 2556	2200			
		1, 2, 3, 17 A, K	1039–2556	12–2200			
napolis 19·0 kc/s	June, 1956				1, 6	5823	75
	Aug., 1956				1, 12	5823, 5940	206
napolis 15·5 kc/s	Dec., 1957	1, 4, 5, 13	5815–5823	31–293			
	Jan., 1958	18, I	5430, 4450	900			
		9, 18, 19	5430–5593	182–800			
		9, 17, 19	5433–5593	182–630			
		1, 9, 17, 18, 19, I	4450–5593	173–1420			
	Feb.–Mar., 1958				1, 14	5823, 5760	286
gby 16·0 kc/s ..	June–July, 1957	J	1280				
ggion 19·6 kc/s..	Mar., 1958	A	2060				

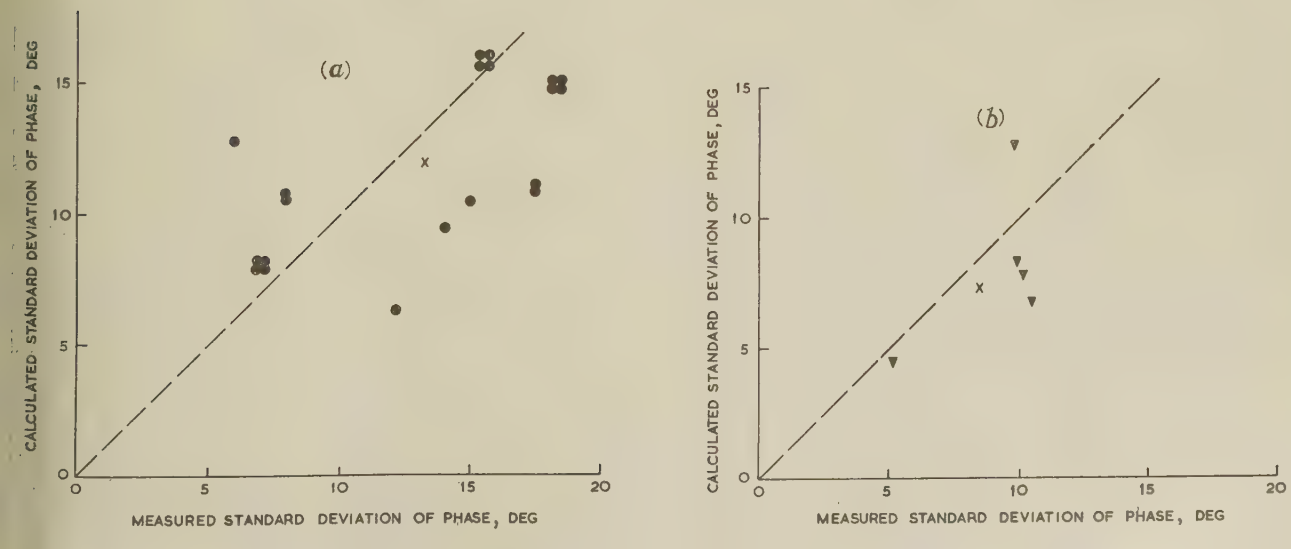


Fig. 9.—Relative phase variations, measured and calculated from amplitude measurements.

(*a*) Range 1000 km, $f = 17{\cdot}25$ kc/s. (*b*) Range 5800 km, $f = 15{\cdot}5$ kc/s.

● $2\frac{1}{2}$ hours observations.
▼ $7\frac{1}{2}$ hours observations.
× Grand average.
—— Random signal.

Tests have been carried out using the observations on Varberg made over the Hemsby–Slough and Lower Whitley–Slough base-lines and on Annapolis over the Carnaby–Slough base-line. The analyses were confined to the night-time period which was clear of the systematic effects associated with sunset and sunrise; the day-time fluctuations were generally very small and not easily separated from the diurnal trends or observational errors. These and the other analyses of amplitude measurements were made on an electronic computer using the values taken from the records at 5 min intervals. Fig. 9 shows the comparison between the measured and the calculated relative phase variations together with the expected theoretical relation. The length of the records used is indicated by the grouping of the points; e.g. in Fig. 9(*a*) a single point indicates the value obtained from 2·5 hours observations and a group of four points that from 10 hours observations. It is noticeable that the longer records give values nearer to the random signal line.

(7.4) Deduced Random Phase Fluctuations

Random phase fluctuations, both absolute and relative, for various ranges and receiver separations have been deduced from the majority of the night-time amplitude measurements.

In Table 7 are given the mean standard deviations of the absolute fluctuations for a number of ranges and for three seasons of the year. The number of periods of observation involved in the calculation of the mean value and the extreme values of standard deviations are also given. The majority of the data was obtained from observations on a frequency of 17·2 kc/s, but data obtained on other frequencies (15·5–19·6 kc/s) are also included.

Although the data are not sufficiently complete to draw any definite conclusions about systematic variations, it appears that there is no significant difference due to season, but there are indications from the other winter data that the fluctuations least at ranges of about 1 500 km and greatest between 2 50 4 500 km. The occurrence of the minimum is also supp by the data for the other seasons. Similar data on the re phase fluctuations for various ranges, seasons, receiver se tions and base-line orientations are given in Table 8.

Again there is no significant difference between summer winter. There is, however, a noticeable change in the ampli of the fluctuations with change of receiver separation; th particularly evident in the results at 5 820 km range in wh

Table 7

STANDARD DEVIATION OF ABSOLUTE PHASE FOR VAR RANGES AND SEASONS COMPUTED FROM AMPLI MEASUREMENTS

Range	Standard deviation of phase					
	Spring		Summer		Winter	
	Mean	Max/min	Mean	Max/min	Mean	Max
km	deg	deg	deg	deg	deg	de
850	10 (3)	15/6	9 (11)	14/4	9 (7)	10
1 000	10 (7)	12/6	8 (30)	14/3	9 (78)	1
1 300	5 (6)	8/3	8 (8)	17/5	6 (8)	
2 100	—	—	3 (4)	5/2	9 (5)	1
2 400	12 (3)	15/9	—	—	11 (6)	1
2 600	—	—	—	—	17 (8)	2
4 500	—	—	—	—	15 (9)	2
5 500	—	—	—	—	11 (60)	2
5 800	—	—	—	—	8 (72)	1

Number of records are shown in brackets.
Max/min signifies maximum and minimum values out of all records at the range.

Table 8

STANDARD DEVIATION OF RELATIVE PHASE FOR VARIOUS RANGES AND SITE SEPARATIONS COMPUTED FROM AMPLITUDE MEASUREMENTS

Mean range	Separation	Base-line orientation	Summer			Winter		
			Standard deviation of relative phase		Standard deviation of absolute phase $\times\sqrt{2}$	Standard deviation of relative phase		Standard deviation of absolute phase $\times\sqrt{2}$
			Mean	Max/Min		Mean	Max/Min	
km	km							
950	206	Radial	10(7)	11/8	12	—	—	—
	275	Oblique	—	—	12	9(6)	12/7	13
	284	Radial	16(2)	18/3	12	—	—	—
1 040	240	Trans	10(6)	13/6	10	12(10)	16/6	13
	555	Trans	10·5(7)	19/6	10	11·5(7)	16/7	13
1 160	304	Oblique	11(6)	13/7	10	—	—	—
2 464	2 200	Trans	—	—	—	21(5)	33/8	20
4 940	900	Radial	—	—	—	20(9)	34/11	20
5 820	50	Trans	—	—	—	4(12)	6/2	13
	212	Trans	—	—	—	10(9)	16/5	13
	293	Trans	—	—	—	11(10)	16/5	13
5 510	490	Trans				13(7)	13/6	17
	630	Trans				11(14)	21/3	17
	650	Trans				15(6)	30/6	17
	800	Trans				9(16)	23/4	17

Trans: Base-line transverse to direction of propagation.
Radial: Base-line in direction of propagation.
Oblique: Base-line oblique to direction of propagation.

tion 50 km is included. If the separation is sufficient for o be no correlation of the fluctuations on the two paths heoretically, the standard deviation of the relative fluctua- should be equal to the square root of the sum of the s of the standard deviations of the absolute fluctuations two paths. On the assumption that the standard devia- of the absolute fluctuations are equal on the two paths eoretical standard deviations of the relative fluctuations o correlation have been computed as shown in the Table. majority of cases the theoretical values are of the same as the measured ones, indicating that there is negligible ation for separations of about 250 km or more. The rements at a range of 5 820 km, however, include separa- below 250 km, and these show a significant decrease in ation amplitude at the smaller separations. This point is with more fully in the next Section.

(7.5) Correlation over Adjacent Paths

value of the correlation coefficient has been obtained he amplitude measurements, using an electronic computer, pairs of stations at which simultaneous observations were Fig. 10 shows the mean coefficient plotted against

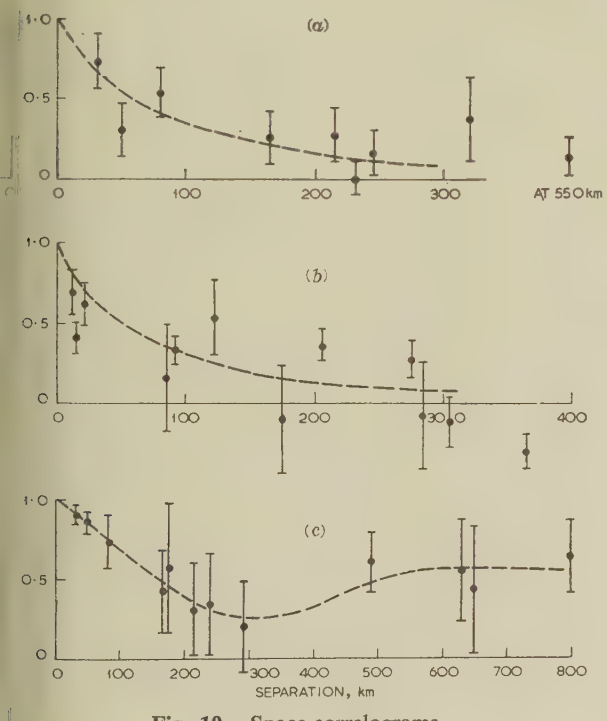


Fig. 10.—Space correlograms.

Range 1000 km. Base-line transverse to direction of propagation.
Range 1000 km. Base-line along direction of propagation.
Range 5800 km. Base-line transverse to direction of propagation.
Vertical lines through points show standard deviations about mean values.

ration for ranges of 1 000 and 5 800 km. Although the er of the points and the standard deviations, as shown by vertical line, are large, it is evident that the correlation mes low at about 200 km separation for 1 000 km range for directions of base-line orientation and at about 300 km ration for 5 800 km range. The values obtained at a range 800 km for separations of 500–800 km, however, show an ase of correlation coefficient to about 0·5. The measure- ts at the 1 000 km range with a separation of 550 km 10(*a*)] indicated that no such increase occurs at the shorter es, and the correlation between measurements at sites at es of 1 040 and 1 404 km along the same radial gave a coefficient of −0·3 [Fig. 10(*b*)]. The reason for this rise in correlation coefficient at the greater separations and ranges is not evident, but a statistical test of its significance shows that, in spite of the large standard deviations involved, such an increase is unlikely to arise by chance. Thus, although we may conclude that for separations of about 250 km and up to at least 550 km there is negligible correlation at ranges of the order of 1 000 km, no similar conclusions can be made at the greater range, although it seems unlikely that the coefficient will rise to near unity.

(8) CONCLUSIONS

The work described was mainly of an exploratory nature in that the phase measurements covered only a limited range of transmitter and receiver configurations and ionospheric conditions, although many additional data on the random phase variations have been obtained from the amplitude measurements. The information collected does, however, give an indication of the effects of the ionosphere on certain aspects of the operation of a phase-comparison type of navigational aid.

The measurements made at the short ranges (100 km) have little practical application to the navigational problem, apart from the checking of the operation of the measuring equipment, since it was only the horizontally polarized component of the ionospheric wave which was selected for examination. The results do show, however, that waves reflected from the ionosphere at points to within about 40 km of each other undergo very similar phase changes, both from day to night and in the short-period fluctuations. This high correlation also holds for separations in the ionosphere of the order of 100 km, except at times of marked magnetic activity. They also show that, as at the longer distances, the phase fluctuations are greater at night than in the day-time. The main value of these short-distance observations lies in their application to the study of ionospheric irregularities which will be dealt with in another paper.

The measurements which are of particular value to the study of navigational aids are those made at medium and long ranges, since it is at such distances that the instability of the ionospheric wave has the greatest influence on the phase of the resultant signal. In summarizing the results of these measurements the diurnal changes and the random fluctuations will be considered separately, the latter including both phase and amplitude measurements.

At ranges of the order of 1 000 km and with receiver separations of 200–250 km, the increases of mean relative phase* from day to night is shown to be of the order of 30° when the receiver base-line is in the direction of propagation and about half this value or less when the base-line is transverse (Table 2). These diurnal changes vary considerably from one day to the next and appear to be independent of season. The largest overall change of phase recorded during a 24-hour period was 72°.

For similar receiver spacing (200–300 km) at ranges of the order of 6 000 km, the change of mean relative phase is appreciably less than at the shorter range. The changes between the night, sunrise, day and sunset periods are of the order of 5° when the receiver base-line is transverse to the direction of propagation (Table 4). Generally, they show a decrease in relative phase from day to night, though the intermediate sunset value may be greater or less than the night one. The overall phase change for the 24-hour period is about half that observed at the shorter range, i.e. 35°.

* Throughout the paper the measured phase values have been quoted in terms of degrees at the particular frequency of measurement. The range of frequencies involved is relatively small (15·5–19·0 kc/s), so that direct comparisons between measurements on different frequencies can be made and any frequency effects which may be present can be ignored. In any case, the method of obtaining the random fluctuations of phase from amplitude measurements assumes that these are independent of frequency.

A general conclusion to be obtained is that, on adjacent paths of comparable length where the mid-path separation is too small (about 100 km) for the relative phase to be significantly affected by the general diurnal movement of the ionosphere, there are irregularities of sufficient size to cause appreciable phase differences. Measurements of the absolute phase changes of 16 kc/s signals over a 6 000 km path (Rugby–Cambridge, Mass.) made by Pierce show diurnal changes of 200°–230°, which are to be compared with the relative changes of about 35° observed by us over a similar transatlantic path on a frequency of 15·5 kc/s. No similar measurements of absolute phase over 1 000 km paths have been made, but assuming that the ionosphere can be represented at very low frequencies by a sharply bounded layer having a reflection coefficient between 0·4 and 0·8 and a diurnal change of effective height from 70 km to 85 km, an absolute phase change of the order of 120° can be expected on a frequency of 17 kc/s. The corresponding relative change observed had an average value of 50° (see Table 2).

The most definite conclusion obtained on the behaviour of the random fluctuations of relative phase is that their amplitude is greater at night than in the day-time by a factor of between 2 and 3 and that this characteristic is independent of range, base-line length or orientation, and season. At ranges of both 1 000 and 6 000 km the average standard deviation of the fluctuations for receiver separations of 200–250 km is about 4° in the day-time with no significant change with season or base-line orientation. At night the standard deviation shows a difference with range, being 11·6° at 1 000 km and 9·4° at 6 000 km.

It has been shown that the magnitude of the phase fluctuations can be deduced from measurements of the amplitude fluctuations. Thus it has been possible to extend the study of the phase fluctuations to greater ranges and receiver spacings by night-time amplitude observations, and to obtain a general picture of the variation of both the absolute and relative phase fluctuations with range, path separation and season (see Tables 6 and 7). In general, the standard deviations of the absolute night-time fluctuations show a minimum of the order of 5° at ranges of about 1 500 km and a peak of the order of 16° in the region 3 000–4 000 km with no significant difference due to season. The analysis of the relative phase values gives standard deviations between 9° and 22° for receiver separations of 200 km or more, but the scope of the observations is not sufficient to show any general trends with range.

An extensive study of the correlation between the phase fluctuations over adjacent paths has shown that at 1 000 km range the correlation falls to a negligible value ($<0·2$) when the receivers are separated by 200 km or more. At 6 000 km range this low correlation is reached at a separation of 300 km, but there is a definite indication that the coefficient rises to about 0·5 at separations between 500 and 800 km. However, since a rise in correlation coefficient leads to a decrease in the amplitude of the fluctuations, we may conclude, in general, that for ranges up to 6 000 km the random relative phase fluctuations observed with receiver separations of about 250 km will not increase in magnitude as the separation is increased. Thus the values given in Tables 3, 5 and 6 may be used generally in assessing the random errors of a system with larger base-li The application of these and other data to the estimation of probable errors of a long-range navigational system will be cussed in a further paper.

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The carrying out of the investigation involved the assista of a number of the authors' colleagues, in particular tha Mr. C. S. Fowler, who was responsible for a major part of development of the equipment and of the organization of observations.

The work described above was undertaken as part of programme of the Radio Research Board. The paper published by permission of the Director of Radio Researc the Department of Scientific and Industrial Research.

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THIRD-HARMONIC TUNING OF E.H.T. TRANSFORMERS

By E. M. CHERRY, M.Sc., Graduate.

(*The paper was first received* 11*th March, and in revised form* 12*th September,* 1960.)

SUMMARY

letailed analysis of third-harmonic tuning of e.h.t. transformers n, using a simple, approximate equivalent circuit. It is shown f the leakage inductance is tuned to about the 2·8th harmonic flyback pulse by the stray capacitance,

a) Ringing of the e.h.t. winding during the scan is eliminated.

b) The e.h.t. voltage pulse is maximized, the increase being 55%.

c) The voltage pulse applied to the driver tubes may be increased lecreased; the maximum possible reduction is 25%.

heoretical transformer should therefore have a pulse step-up of 0·75, i.e. 2·07 times the turns ratio; practical transformers give a step-up of more than 1·5 times the turns ratio. The nation is in the approximate nature of the equivalent circuit used e analysis.

outline of an analysis using a complete equivalent circuit is

1) NATURE AND HISTORY OF THIRD-HARMONIC TUNING

e line output transformer of a television receiver is prone nging effects, which are due to the shock-excited oscillation e resonant circuits formed by the leakage inductances of ransformer with the stray capacitances.

e main windings of a line output transformer, i.e. all the ings except the e.h.t. secondary, are quite closely coupled; eakage inductances are small, so that ringing is at a fre- cy of the order of 1 Mc/s and the oscillating energy is small. ing in the main winding does not pose any serious problems. ever, the e.h.t. winding is necessarily constructed as a fairly ow overwind; the coupling between it and the main winding erefore relatively loose. The combination of large leakage ctance, high voltage and relatively large stray capacitance give rise to oscillations of quite high energy at a frequency few hundred kilocycles per second.

inging of the e.h.t. transformer during the scan is most sirable, as it produces an oscillatory component of current e boost diode. Such an oscillatory current modulates the ge drop across the diode, causing a 'wrinkling' non-linearity e scan. In extreme cases, the diode peak current may be bled, and/or the diode may become cut off during the scan, ing a gross non-linearity as shown in Fig. 1.

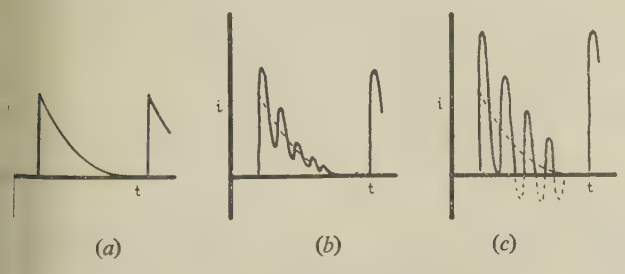


Fig. 1.—Boost-diode current.
(*a*) Without ringing. (*b*) Moderate ringing. (*c*) Serious ringing.

itten contributions on papers published without being read at meetings are d for consideration with a view to publication.
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Third-harmonic tuning of the e.h.t. winding provides a simple means for preventing ringing during the scan. The exact history of third-harmonic tuning is difficult to trace, as very few significant papers have been published. The brevity of Section 9 and the nature of the references included are indicative of the dearth of worthwhile information.

Torsch claims to have done some work on third-harmonic tuning in the late 1940's; judging from an article published in 1953, his work was of an empirical nature only.[1] It was well known that ringing of the e.h.t. winding modified the shape of the flyback pulse at the boost diode and output pentode. The modification is the addition of a sinusoidal voltage at the ringing frequency to the main flyback pulse (Fig. 2). Torsch's contribu-

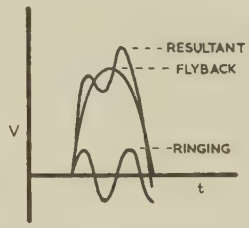


Fig. 2.—Effect of ringing on driver-tube voltage pulse.

tion was to tune the ringing to about the third harmonic of the main flyback pulse; the resultant pulse then has a flattened top and is of reduced peak amplitude, as shown in Fig. 3.

In the early 1950's, papers appeared in Continental journals

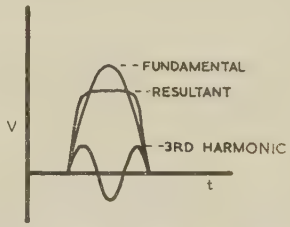


Fig. 3.—Driver-tube voltage pulse with ringing tuned to third harmonic.

which claimed that third-harmonic tuning increased the e.h.t. voltage. The ringing voltage appears at the e.h.t. terminal of the transformer 180° out of phase with that across the main winding, as shown in Fig. 4. However, it was also stated that

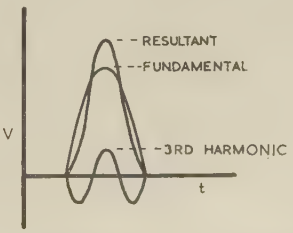


Fig. 4.—E.H.T. voltage pulse, with ringing tuned to third harmonic.

exact third-harmonic tuning produced intolerable ringing during the scan. The suggested remedy was to tune the ringing to a frequency about 10% lower than the third harmonic (i.e. the 2·7th harmonic*) at which the ringing was much reduced. It was claimed that this mistuning did not significantly affect the e.h.t. or main-winding voltage pulses.[2]

The only really significant paper on third-harmonic tuning of which the author is aware[3] was published in 1951. The claim is that, for 2·8th harmonic tuning, ringing of the e.h.t. winding will cease and that the e.h.t. voltage will be increased by 55%. There is no mention of the voltage pulse across the main winding, nor is it claimed that the 55% increase in the e.h.t. voltage is as large as possible. The analysis leading to the claims made in this Patent has not been published.

From 1955, articles have appeared which state that 2·7th harmonic tuning, which gives no ringing, also maximizes the e.h.t. pulse and minimizes the pulse on the driver tubes, i.e. the pulse across the main winding.[4] The author's conclusions differ from these stated results in that, while tuning for zero ringing automatically maximizes the e.h.t. voltage, it does not necessarily minimize the driver-tube voltage pulse.

The most recent papers are more concerned with the voltage pulses than with ringing.[5] There are probably two reasons for this:

(*a*) Slight ringing is tolerable because modern boost diodes have a very low resistance; the ringing current produces only a small modulation of the voltage drop across the diode.

(*b*) The advent of 90° and 110° deflection has resulted in very large pulses on the driver tubes; any means for reducing these pulses is a significant contribution.

The tuning is usually referred to as 'third-harmonic' rather than '2·7th harmonic'.

(2) EQUIVALENT CIRCUIT

A lossless line time-base may be represented as in Fig. 5. The battery and switch simulate the boost capacitor, h.t. supply and driver tubes of a practical circuit.

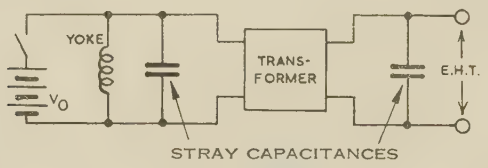


Fig. 5.—Equivalent circuit for line time-base and e.h.t. generator.

The switch is closed, applying a constant voltage, V_0, to the yoke; the yoke current increases linearly with time. As the current reaches I_0, the switch is opened and the resonant circuit formed by the yoke and stray capacitances executes slightly more than one half cycle of oscillation (the flyback). At the end of the flyback, the yoke current is $-I_0$ and the voltage across the yoke is V_0. The switch is closed, and the yoke current flows into the battery, rising linearly from $-I_0$ to zero. If the switch is left closed, the whole process may be repeated and a sawtooth yoke current is obtained. The transformer across the yoke steps up the flyback voltage pulse; the enlarged pulse is rectified to give the e.h.t. voltage.

The problem is to find the condition for the parasitic oscillation, which is shock excited in a practical transformer, to cease at the end of the flyback.

* '2·7th harmonic' is a descriptive, but rather unscientific term, which is in common usage in popular and semi-scientific literature. It is used throughout the paper because it is concise and self-explanatory.

(2.1) Transformer T-Equivalent

The transformer is replaced by its T-equivalent circuit (Fig Two components are neglected to simplify the initial anal

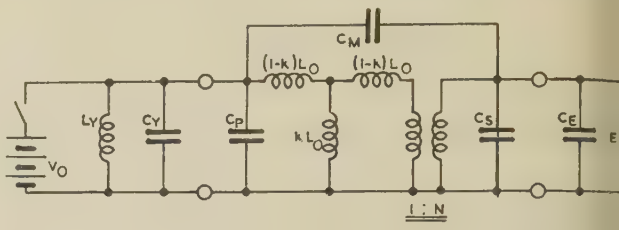


Fig. 6.—Line time-base with transformer replaced by T-equival circuit.

the primary inductance, kL_0, is assumed to be infinite an mutual capacitance, C_M, is assumed to be zero. These ap mations do not introduce gross errors; both components m partially replaced by modifying the values of others.

Next, all components are transferred to the primary of ideal transformer, and any series or parallel componen combined, as shown in Fig. 7. The ideal transformer pla

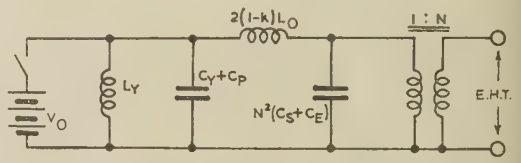


Fig. 7.—Approximate equivalent circuit for line time-base.

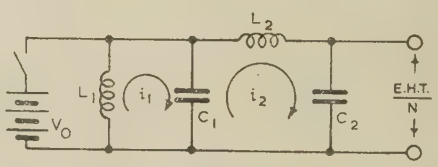


Fig. 8.—Equivalent circuit used in simple analysis.

further part in the analysis and is omitted; the component renamed, giving the final equivalent circuit of Fig. 8.

(2.2) Network Equations

(2.2.1) Initial-Condition Generators.

At the start of the flyback, when the switch is opened conditions are

$$\left.\begin{aligned} i_1 &= I_0 \\ i_2 &= 0 \\ V_{C1} &= V_{C2} = V_0 \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad .$$

and I_0 is given by

$$I_0 = -\frac{V_0}{L_1}\frac{T_s}{2} \quad . \quad . \quad . \quad . \quad . \quad .$$

if the scanning time is T_s. Four generators are placed i equivalent circuit to set up these initial currents and vol (Fig. 9). The required generators are:

Generator 1.—Produces I_0 in L_1, and therefore an im $I_0L_1 = -\frac{1}{2}V_0T_s$, which has a transform $-\frac{1}{2}V_0T_s$.

Generator 2.—Produces the voltage on C_1, and theref step V_0, which has a transform V_0/s.

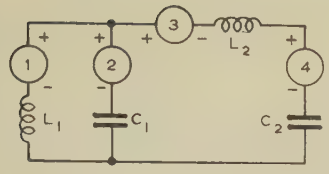


Fig. 9.—Initial-condition generators for simple analysis.
Generator 1: $-\frac{1}{2}V_0T_s$.
Generator 2: V_0/s.
Generator 3: zero.
Generator 4: V_0/s.

erator 3.—There is no current in L_2 and therefore generator ro.

erator 4.—Produces the voltage on C_2, and therefore a $_0$, which has a transform V_0/s.

is the network equations are

$$\begin{bmatrix} + \frac{1}{sC_1} & - \frac{1}{sC_1} \\ - \frac{1}{sC_1} & sL_2 + \frac{1}{s}\left(\frac{1}{C_1} + \frac{1}{C_2}\right) \end{bmatrix} \begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = - V_0 \begin{bmatrix} \frac{1}{s} + \frac{T_s}{2} \\ 0 \end{bmatrix} \quad . \quad . \quad . \quad . \quad (3)$$

Evaluation of i_1 and i_2.

solution of eqn. (3) is

$$i_1 = - \frac{V_0}{L_1}\left(\frac{sT_s}{2} + 1\right)\frac{s^2 + N_0}{s^4 + s^2D_2 + D_0} \quad . \quad . \quad (4a)$$

$$i_2 = - \frac{V_0}{L_1}\left(\frac{sT_s}{2} + 1\right)\frac{n_0}{s^4 + s^2D_2 + D_0} \quad . \quad . \quad (4b)$$

$$\left.\begin{aligned} D_0 &= \frac{1}{L_1L_2C_1C_2} \\ D_2 &= \frac{1}{L_1C_1} + \frac{1}{L_2C_1} + \frac{1}{L_2C_2} \\ N_0 &= \frac{1}{L_2C_1} + \frac{1}{L_2C_2} \\ n_0 &= \frac{1}{L_2C_1} \end{aligned}\right\} \quad . \quad . \quad . \quad (5)$$

ow put

$$s^4 + s^2D_2 + D_0 \equiv (s^2 + \omega_1^2)(s^2 + \omega_2^2) \quad . \quad . \quad (6)$$

e ω_1 and ω_2, which are real, are the angular frequencies of oles of the network. Eqn. (4) then becomes

$$i_1 = - \frac{V_0}{L_1}\left(\frac{sT_s}{2} + 1\right)\frac{s^2 + N_0}{(s^2 + \omega_1^2)(s^2 + \omega_2^2)} \quad . \quad . \quad (7a)$$

$$i_2 = - \frac{V_0}{L_1}\left(\frac{sT_s}{2} + 1\right)\frac{n_0}{(s^2 + \omega_1^2)(s^2 + \omega_2^2)} \quad . \quad . \quad (7b)$$

ext put

$$\left.\begin{aligned} \frac{s^2 + N_0}{(s^2 + \omega_1^2)(s^2 + \omega_2^2)} &\equiv \frac{A}{s^2 + \omega_1^2} + \frac{B}{s^2 + \omega_2^2} \\ \frac{n_0}{(s^2 + \omega_1^2)(s^2 + \omega_2^2)} &\equiv \frac{a}{s^2 + \omega_1^2} + \frac{b}{s^2 + \omega_2^2} \end{aligned}\right\} \quad . \quad (8)$$

e A, B, a and b are the residues at the poles of s^2. Eqn. (7) then be expressed in the form

$$i_1 = - \frac{V_0}{L_1}\frac{T_s}{2}\left[A\left(\frac{s}{s^2 + \omega_1^2} + \frac{2}{\omega_1T_s}\frac{\omega_1}{s^2 + \omega_1^2}\right) + B\left(\frac{s}{s^2 + \omega_2^2} + \frac{2}{\omega_2T_s}\frac{\omega_2}{s^2 + \omega_2^2}\right)\right] \quad (9a)$$

$$i_2 = - \frac{V_0}{L_1}\frac{T_s}{2}\left[a\left(\frac{s}{s^2 + \omega_1^2} + \frac{2}{\omega_1T_s}\frac{\omega_1}{s^2 + \omega_1^2}\right) + b\left(\frac{s}{s^2 + \omega_2^2} + \frac{2}{\omega_2T_s}\frac{\omega_2}{s^2 + \omega_2^2}\right)\right] \quad (9b)$$

ω_1, ω_2, A, B, a and b may be found from eqns. (4)–(8):

$$\left.\begin{aligned} \omega_1^2 + \omega_2^2 &= D_2 \\ \omega_1^2\omega_2^2 &= D_0 \\ A + B &= 1 \\ A\omega_2^2 + B\omega_1^2 &= N_0 \\ a + b &= 0 \\ a\omega_2^2 + b\omega_1^2 &= n_0 \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad (10)$$

Now take the inverse transform of eqn. (9), giving

$$i_1 = - \frac{V_0}{L_1}\frac{T_s}{2}\left[A\left(\cos\omega_1t + \frac{2}{\omega_1T_s}\sin\omega_1t\right) + B\left(\cos\omega_2t + \frac{2}{\omega_2T_s}\sin\omega_2t\right)\right] \quad (11a)$$

$$i_2 = - \frac{V_0}{L_1}\frac{T_s}{2}\left[a\left(\cos\omega_1t + \frac{2}{\omega_1T_s}\sin\omega_1t\right) + b\left(\cos\omega_2t + \frac{2}{\omega_2T_s}\sin\omega_2t\right)\right] \quad (11b)$$

In practice, both $2/\omega_1T_s$ and $2/\omega_2T_s$ are very much less than unity, so to better than 1% we may make the approximations

$$i_1 \simeq - \frac{V_0}{L_1}\frac{T_s}{2}\left[A\cos\left(\omega_1t - \frac{2}{\omega_1T_s}\right) + B\cos\left(\omega_2t - \frac{2}{\omega_2T_s}\right)\right] \quad . \quad . \quad . \quad . \quad (12a)$$

$$i_2 \simeq - \frac{L_0}{V_1}\frac{T_s}{2}\left[a\cos\left(\omega_1t - \frac{2}{\omega_1T_s}\right) + b\cos\left(\omega_2t - \frac{2}{\omega_2T_s}\right)\right] \quad . \quad . \quad . \quad . \quad (12b)$$

Thus, both i_1 and i_2 have components at frequencies ω_1 and ω_2. The most important relation between the coefficients is, from eqn. (10),

$$a = - b \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (13)$$

The phase angles of the ω_1 and ω_2 terms are the same for the two equations, namely $-2/\omega_1T_s$ and $-2/\omega_2T_s$.

(3) CONDITIONS FOR RINGING TO CEASE

The flyback in a practical line time-base finishes when the boost diode begins to conduct. This is simulated in the equivalent circuit by closing the switch at an instant when

$$V_{C1} = - L_1\frac{di_1}{dt} = V_0 \quad . \quad . \quad . \quad . \quad (14)$$

If, at this instant, $di_2/dt = 0$, then

$$V_{L2} = - L_2\frac{di_2}{dt} = 0$$

and therefore $V_{C1} = V_{C2} = V_0$.

Further, if at this instant $i_2 = 0$, then, since $V_{C1} = V_{C2}$, there is no tendency for i_2 to change from zero, i.e. i_2 will remain at zero during the scan and there will be no ringing.

Thus the conditions for ringing to cease are

$$i_2(T_f) = 0 \qquad \frac{di_2(T_f)}{dt} = 0 \quad . \quad . \quad . \quad . \quad (15)$$

where T_f is the flyback time.

(4) SOLUTION OF EQUATIONS

In theory, eqns. (12*a*) and (14) may be solved to give the flyback time in terms of the parameters L_1, L_2, C_1 and C_2. Assignment of a practical value to T_f—typically 10 microsec—gives one equation for L_1, L_2, C_1 and C_2. Eqns. (12*b*) and (15) may be manipulated to give two further equations in these parameters. If a numerical value is given to any one of the four parameters, the three others may be found from these three equations, but such a solution is exceedingly tedious.

(4.1) Practical Solution

Fortunately, a commonsense argument leads to a workable solution.

The equivalent circuit is lossless, so it is possible to have a completely symmetrical flyback. This implies that

$$\left.\begin{aligned} i_1(t) &= -\,i_1(T_f - t) \\ i_2(t) &= -\,i_2(T_f - t) \\ \frac{di_1(t)}{dt} &= +\frac{di_1(T_f - t)}{dt} \\ \frac{di_2(t)}{dt} &= +\frac{di_2(T_f - t)}{dt} \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad (16)$$

But the initial conditions are, from eqn. (1),

$$i_2(0) = 0 \qquad \frac{di_2(0)}{dt} = 0 \quad . \quad . \quad . \quad . \quad . \quad (17)$$

Therefore, if the flyback is symmetrical,

$$i_2(T_f) = 0 \qquad \frac{di_2(T_f)}{dt} = 0$$

which satisfy eqn. (15), i.e. the conditions for ringing to cease.

(4.1.1) Symmetrical Flyback.

For a symmetrical flyback, $i_2(T_f) = 0$. Therefore, from eqns. (12*b*) and (13), it immediately follows that

$$\cos\left(\omega_1 T_f - \frac{2}{\omega_1 T_s}\right) = \cos\left(\omega_2 T_f - \frac{2}{\omega_2 T_s}\right) \quad . \quad (18)$$

By appeal to the symmetry of the flyback, as in Fig. 10, the flyback time may be evaluated:

$$\omega_1 T_f = \pi + 2\left(\frac{2}{\omega_1 T_s}\right)$$

which gives

$$T_f = \frac{1}{\omega_1}\left(\pi + \frac{4}{\omega_1 T_s}\right) \quad . \quad . \quad . \quad . \quad (19)$$

Substitution from eqn. (19) into eqn. (18) gives

$$\cos\left(\pi + \frac{2}{\omega_1 T_s}\right) = \cos\left[\frac{\omega_2}{\omega_1}\left(\pi + \frac{4}{\omega_1 T_s}\right) - \frac{2}{\omega_2 T_s}\right] \quad (20)$$

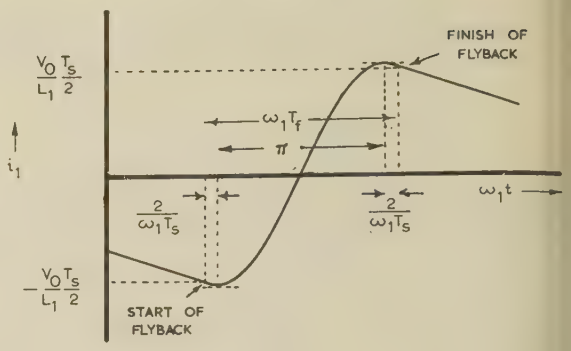


Fig. 10.—Evaluation of flyback time.

(4.1.2) First Solution.

The first non-trivial solution of eqn. (20) is

$$\left(\pi + \frac{2}{\omega_1 T_s}\right) + 2\pi = \frac{\omega_2}{\omega_1}\left(\pi + \frac{4}{\omega_1 T_s}\right) - \frac{2}{\omega_2 T_s}$$

i.e.

$$3\pi + \frac{2}{\omega_1 T_s} = \frac{\omega_2}{\omega_1}\left(\pi + \frac{4}{\omega_1 T_s}\right) - \frac{2}{\omega_2 T_s}$$

Therefore

$$\frac{\omega_2}{\omega_1} = 3 + \frac{2/\omega_2 T_s - 10/\omega_1 T_s}{\pi + 4/\omega_1 T_s}$$

and neglecting small terms,

$$\frac{\omega_2}{\omega_1} \simeq 3 - \frac{10}{\pi \omega_1 T_s} \quad . \quad . \quad . \quad . \quad .$$

Very approximately, from eqn. (19),

$$\omega_1 = \frac{\pi}{T_f} \quad . \quad . \quad . \quad . \quad . \quad .$$

Substitution into eqn. (21) gives

$$\frac{\omega_2}{\omega_1} \simeq 3\left(1 - \frac{10 T_f}{3\pi^2 T_s}\right) \quad . \quad . \quad . \quad .$$

and if $T_f = 0{\cdot}2T_s$, approximately the ratio of flyback to time in a practical television system,

$$\frac{\omega_2}{\omega_1} = 2{\cdot}80 \quad . \quad . \quad . \quad . \quad . \quad .$$

Knowing the order of the answer, and modifying the app mations accordingly, gives a rather more accurate expressio

$$\frac{\omega_2}{\omega_1} = 3\left(1 - 0{\cdot}30\frac{T_f}{T_s}\right) \quad . \quad . \quad . \quad .$$

(4.1.3) Second Solution.

A second solution of eqn. (20) is

$$\left(\pi + \frac{2}{\omega_1 T_s}\right) + 4\pi = \frac{\omega_2}{\omega_1}\left(\pi + \frac{4}{\omega_1 T_s}\right) - \frac{2}{\omega_2 T_s}$$

which yields

$$\frac{\omega_2}{\omega_1} \simeq 5\left(1 - 0{\cdot}37\frac{T_f}{T_s}\right) \quad . \quad . \quad . \quad .$$

and for $T_f = 0{\cdot}2T_s$,

$$\frac{\omega_2}{\omega_1} = 4{\cdot}63$$

This solution corresponds to a flattening of the e.h.t. pulse an increase in the pulse applied to the driver tubes.[4]

The constants in eqns. (25) and (26) are rather smaller

of Reference 3, and it is concluded that the above analysis erent from that which led to the Patent.

(5) CONSEQUENCES OF SOLUTION

(5.1) Relations between L_1, L_2, C_1 and C_2

$$C_2 = pC_1 \quad L_2 = qL_1 \quad . \quad . \quad . \quad . \quad . \quad (27)$$

(5) and (10) may be solved to give

$$\text{or } \omega_2^2 = \frac{1}{L_1C_1}\,\frac{(pq+p+1) \mp \sqrt{[(pq+p+1)^2 - 4pq]}}{2pq} \quad . \quad . \quad . \quad . \quad (28)$$

nediately follows that

$$\left(\frac{\omega_2}{\omega_1}\right)^2 = \frac{(pq+p+1) + \sqrt{[(pq+p+1)^2 - 4pq]}}{(pq+p+1) - \sqrt{[(pq+p+1)^2 - 4pq]}}$$

may be expressed in the form

$$(pq+p+1)^2 = \left[\frac{(\omega_2/\omega_1)^2 + 1}{\omega_2/\omega_1}\right]^2 pq \quad . \quad . \quad (29)$$

utting $\omega_2/\omega_1 = 2{\cdot}80$ [eqn. (24)] gives

$$(pq+p+1)^2 = 10{\cdot}0pq \quad . \quad . \quad . \quad . \quad (30)$$

her p or q is chosen, the value of the other which gives ct tuning may be found from eqn. (30).

(5.2) Flyback Time

stitution from eqn. (30) into eqn. (28) gives

$$\omega_1^2 = \frac{1}{L_1C_1}\,\frac{\sqrt{10}-\sqrt{6}}{2\sqrt{(pq)}} = \frac{0{\cdot}36}{L_1C_1\sqrt{(pq)}} \quad . \quad . \quad (31a)$$

$$\omega_2^2 = \frac{1}{L_1C_1}\,\frac{\sqrt{10}+\sqrt{6}}{2\sqrt{(pq)}} = \frac{2{\cdot}81}{L_1C_1\sqrt{(pq)}} \quad . \quad . \quad (31b)$$

bstitution from eqn. (31a) into eqn. (22) gives for the fly- time

$$T_f \simeq 5{\cdot}3\sqrt{(L_1C_1)}\sqrt[4]{(pq)} \quad . \quad . \quad . \quad . \quad (32)$$

(5.3) E.H.T. Voltage

e e.h.t. voltage at any instant is given by the indefinite ral

$$V_{C2} = \frac{1}{C_2}\int i_2 dt \quad . \quad . \quad . \quad . \quad . \quad (33)$$

ated at the desired instant; the constant of integration is . Therefore, from eqn. (12b),

$$V_{C2} = -\frac{V_0T_s}{2L_1C_2}\left[\frac{a}{\omega_1}\sin\left(\omega_1 t - \frac{2}{\omega_1T_s}\right) + \frac{b}{\omega_2}\sin\left(\omega_2 t - \frac{2}{\omega_2T_s}\right)\right] \quad (34)$$

follows from symmetry that, at the peak of the flyback,

$$\left.\begin{aligned} t &= \frac{T_f}{2} \\ \omega_1 t - \frac{2}{\omega_1 T_s} &= \frac{\pi}{2} \\ \omega_2 t - \frac{2}{\omega_2 T_s} &= \frac{3\pi}{2} \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad (35)$$

Substitution into eqn. (34) gives

$$\hat{V}_{C2} = -\frac{V_0T_s}{2L_1C_2}\left(\frac{a}{\omega_1} - \frac{b}{\omega_2}\right)$$

and expansion of $a/\omega_1 - b/\omega_2$ leads to

$$\hat{V}_{C2} = -\frac{V_0T_s\omega_1\omega_2}{2L_1C_2(\omega_2-\omega_1)}\left(\frac{a\omega_2^2 + b\omega_1^2}{\omega_1^2\omega_2^2} - \frac{a+b}{\omega_1\omega_2}\right) \quad (36)$$

Substitution from eqn. (10) gives

$$\hat{V}_{C2} = -\frac{V_0T_s\omega_1\omega_2 n_0}{2L_1C_2(\omega_2-\omega_1)D_0}$$

Evaluating from eqn. (5),

$$\hat{V}_{C2} = -\frac{\omega_1\omega_2}{\omega_2-\omega_1}\,\frac{V_0T_s}{2}$$

and using eqns. (2) and (24),

$$\hat{V}_{C2} = -1{\cdot}55\omega_1V_0\frac{T_s}{2} = +1{\cdot}55\omega_1L_1I_0 \quad . \quad . \quad (37)$$

It may be shown that the e.h.t. voltage from a system with the same flyback time and a tightly coupled transformer is

$$-1{\cdot}00\omega_1V_0\frac{T_s}{2} \quad . \quad . \quad . \quad . \quad . \quad (38)$$

Therefore there is a gain of 55% if 2·8th harmonic tuning is employed.

(5.4) Driver-Tube Voltage

The 2·8th harmonic tuning, with its consequent 55% increase in e.h.t. voltage, is assured if eqn. (30) is satisfied. Either p or q may be given any value, within limits. It will now be shown that, for one particular set of values of p and q, the voltage pulse across the yoke, i.e. applied to the driver tubes, is as small as possible.

The driver-tube voltage is given by

$$V_{C1} = V_{L1} = -L_1\frac{di_1}{dt} \quad . \quad . \quad . \quad . \quad (39)$$

which may be evaluated from eqn. (12a) as

$$V_{C1} = -V_0\frac{T_s}{2}A\omega_1\left[\sin\left(\omega_1 t - \frac{2}{\omega_1T_s}\right) + \frac{B\omega_2}{A\omega_1}\sin\left(\omega_2 t - \frac{2}{\omega_2T_s}\right)\right] \quad (40)$$

In contrast to the e.h.t. voltage, V_{C2}, the maximum value of V_{C1} need not occur at $t = \frac{1}{2}T_f$. It may be shown graphically, by plotting $(\cos\theta + K\cos 2{\cdot}8\theta)$ for various values of K, that the maximum value of V_{C1} is as small as possible if

$$K \equiv \frac{B\omega_2}{A\omega_1} \simeq 0{\cdot}2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (41)$$

This condition is very broad. Therefore, since ω_2/ω_1 is of the order of 3, it is sufficient, in order to minimize the pulse applied to the drive tubes, that

$$A \simeq 15B \quad . \quad . \quad . \quad . \quad . \quad . \quad (42)$$

If $B\omega_2/A\omega_1 \leqslant 0{\cdot}2$, the maximum of V_{C1} is at $t = \frac{1}{2}T_f$. If $B\omega_2/A\omega_1 > 0{\cdot}2$, two maxima occur symmetrically about the mid-point. Substitution into eqn. (40) then gives for the minimum driver-tube voltage pulse

$$\hat{V}_{C1} = -0{\cdot}75\omega_1V_0\frac{T_s}{2} \quad . \quad . \quad . \quad . \quad (43)$$

The peak yoke voltage for a system with the same flyback time and a tightly coupled transformer may be shown to be

$$-1{\cdot}00\omega_1 V_0 \frac{T_s}{2} \quad . \quad . \quad . \quad . \quad . \quad (44)$$

Therefore there can be a reduction of up to 25% if 2·8th harmonic tuning is properly employed.

(5.4.1) **Evaluation of p and q.**

From eqns. (5) and (10),

$$A\omega_2^2 + B\omega_1^2 = \frac{1}{L_2C_1} + \frac{1}{L_2C_2}$$

This may be expanded from eqns. (27) and (31) to give

$$\frac{2{\cdot}81A + 0{\cdot}36B}{L_1C_1\sqrt{(pq)}} = \frac{1}{L_1C_1}\left(\frac{p+1}{pq}\right)$$

i.e.
$$2{\cdot}81A + 0{\cdot}36B = \frac{p+1}{\sqrt{(pq)}} \quad . \quad . \quad . \quad . \quad (45)$$

Eqn. (10) gives $A + B = 1$. Using eqn. (42) it then follows that

$$A = \frac{15}{16} \qquad B = \frac{1}{16} \quad . \quad . \quad . \quad . \quad . \quad (46)$$

Substitution into eqn. (45) yields

$$p + 1 = 2{\cdot}65\sqrt{(pq)} \quad . \quad . \quad . \quad . \quad (47)$$

Eqns. (30) and (47) may be solved to give the non-trivial solution

$$p = 0{\cdot}36 \qquad q = 0{\cdot}73 \quad . \quad . \quad . \quad . \quad (48)$$

(5.4.2) **Evaluation of T_f.**

Substitution from eqn. (48) into eqn. (32) gives

$$T_f = 3{\cdot}8\sqrt{(L_1C_1)} \quad . \quad . \quad . \quad . \quad . \quad (49)$$

If L_1, the yoke inductance, is chosen, the optimum values of all other components (referred to the yoke terminals of the transformer) are uniquely determined by the flyback time [eqns. (48) and (49)].

(6) CONCLUSIONS FROM SIMPLE ANALYSIS

If an e.h.t. system can be replaced by the equivalent circuit of Fig. 8, there will be no ringing if the poles of the network are in the ratio ω_2/ω_1 given by eqn. (23). The poles of the network will be in the desired ratio if eqn. (30) is satisfied.

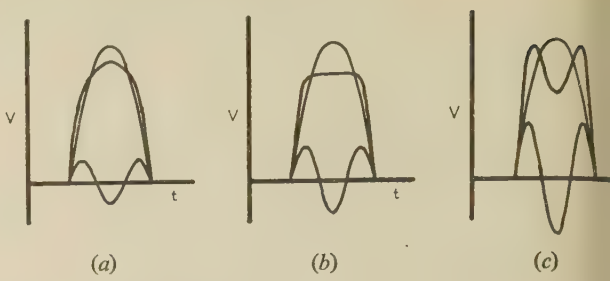
Both the e.h.t. and driver-tube voltages have components at each of the frequencies ω_1 and ω_2. Correct tuning gives a symmetrical flyback, and the phase relations are such that:

(*a*) The components add in the case of e.h.t. waveform; the peak voltage is increased by 55% relative to that from a tightly coupled transformer with the same flyback time.

(*b*) The components subtract in the case of the driver-tube waveform; the voltage halfway through the flyback is reduced, but the maximum (which need not occur halfway through the flyback) may be either increased or decreased.

If $p = 0{\cdot}36$ and $q = 0{\cdot}73$ [eqn. (48)] the maximum driver-tube voltage is as small as possible. The reduction is 25% relative to the peak from a tightly coupled transformer with the same flyback time. With these values of p and q, the flyback time is $T_f = 3{\cdot}8\sqrt{(L_1C_1)}$ [eqn. (49)].

If $p < 0{\cdot}36$ and q is adjusted to give correct tuning, the driver-tube voltage halfway through the flyback is reduced by less than 25% and is the maximum. If $p > 0{\cdot}36$, the driver-tube voltage halfway through the flyback is reduced by more than 25% and two voltage maxima occur symmetrically a this point. This is illustrated in Fig. 11.



(*a*) (*b*) (*c*)

Fig. 11.—Effect of e.h.t. winding capacitance on driver-tube vo
(*a*) $p < 0{\cdot}36$. (*b*) $p = 0{\cdot}36$. (*c*) $p > 0{\cdot}36$.

(6.1) Theory and Practice

The results of the foregoing analysis are confirmed qualitat in practice. Quantitatively, the agreement is fairly good i points except the effective pulse step-up. Theory predi effective step-up of slightly more than twice the turns r i.e. $1{\cdot}55/0{\cdot}75 = 2{\cdot}07$. In practice, the effective ratio is more than 1·5. The explanation is in the approximate n of the equivalent circuit. In particular, no account has taken of losses, and the primary inductance and the m capacitance of the transformer have been neglected. The line of an analysis using a more complete equivalent circ given in Section 7.

The purpose of this more complete analysis is to show the inclusion of kL_0 and C_M decreases the effective pulse ste and further, that the step-up is decreased approximately t value observed in practice. No account is taken of losses is shown in Section 7.7 that there is no general solution equations in the lossy case. Special solutions of the type lined in Section 4 exist for isolated sets of component va but the labour involved in finding such a solution is out proportion to the value of the results which would be obta The whole equivalent circuit is at the best an approxima for example, it is an oversimplification to lump the distrib capacitance of a transformer winding across the ends o winding. If a solution for a lossy transformer were foun would not apply accurately to any practical transformer.

(6.2) Design of Practical Transformers

The design of a practical e.h.t. transformer should be base the above simple analysis, with every effort made to minimiz effects of kL_0 and C_M, i.e. the primary inductance shou large and the mutual capacitance small.

The main winding should be constructed in accordance well-known principles,[6] the primary inductance being as by Friend;[7] with modern core materials, the coefficien coupling can be made very close to unity, so that Fri formula will satisfy the requirement of large primary induct The primary winding capacitance is adjusted to give the req flyback time [eqn. (49)].

The number of turns in the e.h.t. winding is calcul assuming an effective pulse step-up of 1·5 times the turns i.e. assuming that eqn. (37) becomes

$$\hat{V}_{C2} = -1{\cdot}15\omega_1 V_0 \frac{T_s}{2}$$

and eqn. (43) remains as

$$\hat{V}_{C1} = -0{\cdot}75\omega_1 V_0 \frac{T_s}{2}$$

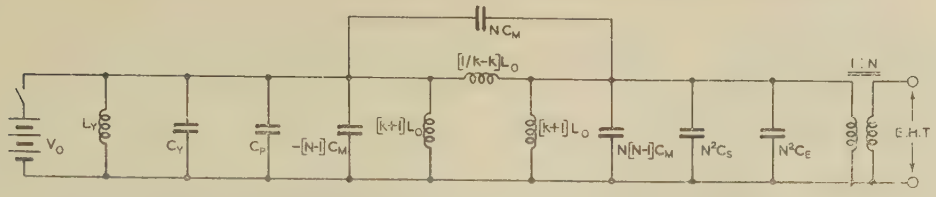


Fig. 12.—Equivalent circuit for line time-base, after manipulation.

ll be suggested by eqns. (66) and (68) respectively. The winding capacitance is calculated from eqn. (48), and a er of transformers are constructed with capacitances ng around this value. These transformers are tuned by ting the coupling between the windings to give minimum in the boost-diode current waveform, i.e. by adjusting q n. (48) to its required value.

e waveforms of the pulses across the main windings of these transformers are examined. It will be found that, as the winding capacitance increases, so the main-winding flyback s change from pointed through flat-topped to double-ped. The transformer which gives the flattest pulse is then ed for e.h.t. voltage under load, and the number of turns e.h.t. winding adjusted as required. If the change is more a few per cent, the entire process should be repeated until isfactory design is obtained. In all cases, transformers d be tuned to give minimum ripple in the boost-diode nt, rather than maximum or minimum voltage pulses.

) Auto-Transformer Connection.

e.h.t. transformer is usually constructed as an auto-trans-er, with the start of the e.h.t. winding returned to the finish e main winding. Any capacitance from the e.h.t. terminal rth, e.g. the e.h.t.-rectifier-tube capacitance, forms part of nutual capacitance and not part of the capacitance across .h.t. winding. Such capacitance should be minimized.

(7) COMPLETE ANALYSIS

(7.1) Equivalent Circuit

a starting-point, consider the lossless equivalent circuit of 6. The equivalent circuit is manipulated as follows:

a) C_S and C_E are moved to the primary side of the ideal nsformer.
b) A star-delta transform is performed on the inductances kL_0 $(1-k)L_0$.
c) C_M is replaced by three capacitances $-(N-1)C_M$, NC_M and $N-1)C_M$.

ie circuit is then as in Fig. 12. Parallel components are bined and the resulting elements are renamed, giving the , complete, lossless equivalent circuit of Fig. 13.

(7.2) Existence of Solution

ie simple analysis is dependent on the network having only sets of conjugate poles. The network for the complete ysis has an additional two elements; its circuit determinant erefore of the sixth order and it is expected that there will be e sets of conjugate poles. However, it is easily demonstrated the driving-point and transfer impedances have two poles . A solution modelled on that of Section 4.1 will therefore pplicable.

ie reactance plot for a parallel LC circuit is shown in 14(*a*). If two such circuits are connected in series, the tance plot has two poles as in Fig. 14(*b*). If a third LC uit is connected in shunt with this series arrangement, the ltant plot still has two poles only, as in Fig. 14(*c*). It is

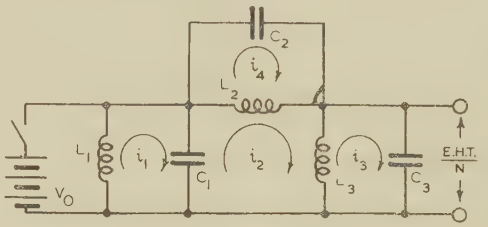


Fig. 13.—Equivalent circuit used in complete analysis.

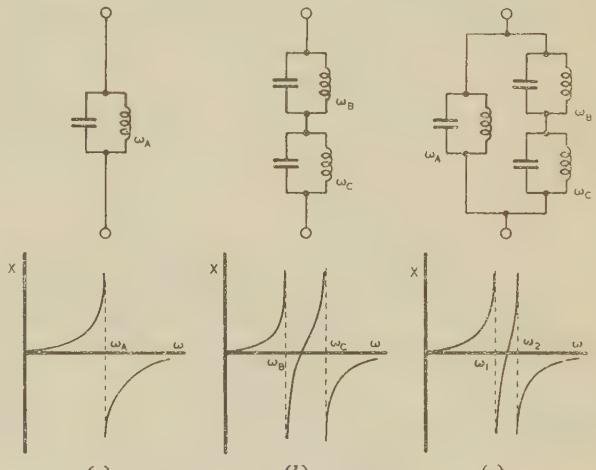


Fig. 14.—Reactance plots for equivalent circuit, showing that there are two pairs of conjugate poles only in the driving-point impedance.

expected, therefore, that there will be a common factor of degree s^2 in the circuit determinants.

(7.3) Network Equations

The required initial conditions are

$$\left.\begin{aligned} i_1 &= I_0 = -\frac{V_0 T_s}{2L_1} \\ i_2 &= \frac{V_0 T_s}{2(L_2+L_3)} \\ i_3 &= i_4 = 0 \\ V_{C1} &= V_0 \\ V_{C2} &= \frac{V_0 L_2}{L_2+L_3} \\ V_{C3} &= \frac{V_0 L_3}{L_2+L_3} \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad (50)$$

If generators with the transforms shown in Fig. 15 are included in the equivalent circuit, these conditions will be simulated.

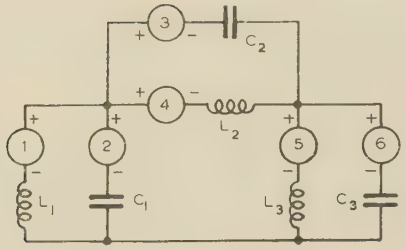


Fig. 15.—Initial-condition generators for complete analysis.

Generator 1: $-\frac{1}{2}V_0T_s$.
Generator 2: V_0/s.
Generator 3: $V_0L_2/(L_2 + L_3)s$
Generator 4: $-V_0L_2T_s/2(L_2 + L_3)$.
Generator 5: $-V_0L_3T_s/2(L_2 + L_3)$.
Generator 6: $V_0L_3/(L_2 + L_3)s$.

The network equations are

$$\begin{bmatrix} sL_1 + \frac{1}{sC_1} & -\frac{1}{sC_1} & 0 & 0 \\ -\frac{1}{sC_1} & sL_2 + sL_3 + \frac{1}{sC_1} & -sL_3 & -sL_2 \\ 0 & -sL_3 & sL_3 + \frac{1}{sC_3} & 0 \\ 0 & -sL_2 & 0 & sL_2 + \frac{1}{sC_2} \end{bmatrix} \begin{bmatrix} i_1 \\ i_2 \\ i_3 \\ i_4 \end{bmatrix}$$

$$= V_0\left(\frac{T_s}{2} + \frac{1}{s}\right)\begin{bmatrix} 1 \\ +1 \\ -\frac{L_3}{L_2 + L_3} \\ -\frac{L_2}{L_2 + L_3} \end{bmatrix} \quad . \quad . \quad . \quad (51)$$

The relevant solutions of eqn. (51) are

$$i_1 = -\frac{V_0}{L_1}\left(\frac{sT_s}{2} + 1\right)\frac{s^2 + N_0}{s^4 + s^2D_2 + D_0} \quad . \quad . \quad (52a)$$

$$i_3 = -\frac{V_0}{L_1}\left(\frac{sT_s}{2} + 1\right)\frac{s^2n_2 + n_0}{s^4 + s^2D_2 + D_0} \quad . \quad . \quad (52b)$$

where

$$\left.\begin{aligned} D_0 &= \frac{1}{C_2C_3 + C_3C_1 + C_1C_2}\left(\frac{1}{L_2L_3} + \frac{1}{L_3L_1} + \frac{1}{L_1L_2}\right) \\ D_2 &= \frac{1}{C_2C_3 + C_3C_1 + C_1C_2}\left[C_1\left(\frac{1}{L_2} + \frac{1}{L_3}\right) + C_2\left(\frac{1}{L_3} + \frac{1}{L_1}\right) + C_3\left(\frac{1}{L_1} + \frac{1}{L_2}\right)\right] \\ N_0 &= \frac{C_1(L_2 + L_3)^2 + C_2L_2^2 + C_3L_3^2}{(C_2C_3 + C_3C_1 + C_1C_2)L_2L_3(L_2 + L_3)} \\ n_0 &= \frac{C_3(L_1 + L_2 + L_3)}{(C_2C_3 + C_3C_1 + C_1C_2)L_2(L_2 + L_3)} \\ n_2 &= \frac{C_2C_3(L_1 + L_2 + L_3)}{(C_2C_3 + C_3C_1 + C_1C_2)(L_2 + L_3)} \end{aligned}\right\} \quad (53)$$

As expected from Section 7.2, a factor $s^2/L_1C_1C_2C_3$ has cancelled in all determinants.

As in the simple analysis, it will follow that

$$i_1 \simeq -\frac{V_0}{L_1}\frac{T_s}{2}\left[A\cos\left(\omega_1 t - \frac{2}{\omega_1 T_s}\right) + B\cos\left(\omega_2 t - \frac{2}{\omega_2 T_s}\right)\right] \quad . \quad . \quad . \quad . \quad (54a)$$

$$i_3 \simeq -\frac{V_0}{L_1}\frac{T_s}{2}\left[a\cos\left(\omega_1 t - \frac{2}{\omega_1 T_s}\right) + b\cos\left(\omega_2 t - \frac{2}{\omega_2 T_s}\right)\right] \quad . \quad . \quad . \quad . \quad (54b)$$

$$\left.\begin{aligned} \omega_1^2 + \omega_2^2 &= D_2 \\ \omega_1^2\omega_2^2 &= D_0 \\ A + B &= 1 \\ A\omega_2^2 + B\omega_1^2 &= N_0 \\ a + b &= n_2 \\ a\omega_2^2 + b\omega_1^2 &= n_0 \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad (55)$$

For ringing to cease, the flyback must be symmetrical; therefore, among other conditions, $i_3(T_f) = 0$. Substitution in eqn. (19) into eqn. (54*b*) then gives

$$a\cos\left[\left(\pi + \frac{4}{\omega_1 T_s}\right) - \frac{2}{\omega_1 T_s}\right] + b\cos\left[\frac{\omega_2}{\omega_1}\left(\pi + \frac{4}{\omega_1 T_s}\right) - \frac{2}{\omega_2 T_s}\right] = 0$$

(7.4) Departure from Simple Analysis

In general, from eqns. (53) and (55),

$$a \neq -b \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (56)$$

so a departure must now be made from the simple analysis.

An initial condition is $i_3(0) = 0$. Therefore, from eqn. (54*b*),

$$a\cos\left(-\frac{2}{\omega_1 T_s}\right) + b\cos\left(-\frac{2}{\omega_2 T_s}\right) \simeq 0$$

i.e.
$$b \simeq -a\frac{\cos(-2/\omega_1 T_s)}{\cos(-2/\omega_2 T_s)} \quad . \quad . \quad . \quad . \quad (57)$$

We now anticipate the result and put

$$\frac{\omega_2}{\omega_1} = 3(1 - \delta) \quad . \quad . \quad . \quad . \quad . \quad . \quad (58)$$

in eqn. (56), giving

$$a\cos\left[\left(\pi + \frac{4}{\omega_1 T_s}\right) - \frac{2}{\omega_1 T_s}\right] - a\frac{\cos(-2/\omega_1 T_s)}{\cos(-2/\omega_2 T_s)}\cos\left[3(1-\delta)\left(\pi + \frac{4}{\omega_1 T_s}\right) - \frac{2}{\omega_2 T_s}\right]$$

which simplifies to

$$\cos\left(-\frac{2}{\omega_2 T_s}\right) = \cos\left[3\pi\delta - \frac{12}{\omega_1 T_s}(1 - \delta) + \frac{2}{\omega_2 T_s}\right]$$

for which a solution is

$$-\frac{2}{\omega_2 T_s} = 3\pi\delta - \frac{12}{\omega_1 T_s}(1 - \delta) + \frac{2}{\omega_2 T_s}$$

Therefore
$$\delta = \frac{4/\omega_1 T_s - 4/3\omega_2 T_s}{\pi + 4/\omega_1 T_s} \simeq \frac{4}{\pi\omega_1 T_s}$$

and using eqn. (22),

$$\delta \simeq \frac{4T_f}{\pi^2 T_s}$$

fore, from eqn. (59),

$$\frac{\omega_2}{\omega_1} = 3\left(1 - \frac{4T_f}{\pi^2 T_s}\right) \quad . \quad . \quad . \quad . \quad . \quad (60)$$

is nearly the same as eqn. (23).

ns. (23) and (60) are different because they have been ed by different approximate methods. Of the two, the ation of eqn. (60) is more general, but that of eqn. (23) is accurate if C_2 is small. The derivation of eqn. (60) involves pproximate equation (58), which is in direct contradiction n. (13), the latter being exact if C_2 is zero.

attempt has been made to produce an expression corre- ling to eqn. (30) of the simple analysis. Such a formula d involve four ratios instead of two, and would be of little

(7.5) E.H.T. Voltage

in the simple analysis [eqn. (36)] the e.h.t. voltage is by

$$\hat{V}_{C3} = -\frac{V_0 T_s \omega_1 \omega_2}{2L_1 C_3(\omega_2 - \omega_1)}\left(\frac{a\omega_2^2 + b\omega_1^2}{\omega_1^2\omega_2^2} - \frac{a+b}{\omega_1\omega_2}\right)$$

titution from eqn. (55) gives

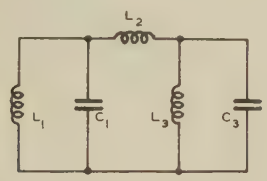
$$\hat{V}_{C3} = -\frac{V_0 T_s \omega_1 \omega_2}{2L_1 C_3(\omega_2 - \omega_1)}\left(\frac{n_0}{D_0} - \frac{n_2}{\sqrt{D_0}}\right) \quad . \quad (61)$$

neral, the constants may be evaluated from eqn. (53).

) **Special Case of $C_2 = 0$.**

C_2 tends to zero, the equivalent circuit becomes as in 16, and eqns. (53) and (61) lead to

$$\hat{V}_{C3} = -\frac{\omega_1\omega_2}{\omega_2 - \omega_1}\frac{V_0 T_s}{2}\left(1 - \frac{L_2}{L_2 + L_3}\right) \quad . \quad (62)$$



16.—A fairly accurate simple equivalent circuit, obtained by neglecting the mutual capacitance, C_M.

2) **Special Case of $L_3 = \infty$.**

L_3 tends to infinity, the equivalent circuit becomes as in 17, and eqns. (53) and (61) lead to

$$\hat{V}_{C3} = -\frac{\omega_1\omega_2}{\omega_2 - \omega_1}\frac{V_0 T_s}{2}\left[1 - \frac{C_2\sqrt{(L_2/L_1)}}{\sqrt{(C_2C_3 + C_3C_1 + C_1C_2)}}\right] \quad (63)$$

$_2$ is small compared with C_1 and C_3,

$$\hat{V}_{C3} \simeq -\frac{\omega_1\omega_2}{\omega_2 - \omega_1}\frac{V_0 T_s}{2}\left[1 - \frac{C_2}{C_1}\sqrt{\left(\frac{L_2/L_1}{C_3/C_1}\right)}\right] \quad (64)$$

ig. 17 is the equivalent circuit most commonly used in ers on third-harmonic tuning. With likely component values, . 16 is a more accurate simple equivalent circuit.

.3) **Approximate General Expression.**

rom eqns. (62) and (64), an approximate general expression the e.h.t. voltage is

$$\hat{V}_{C3} \simeq -\frac{\omega_1\omega_2}{\omega_2 - \omega_1}\frac{V_0 T_s}{2}\left[1 - \frac{L_2}{L_2 + L_3} \cdot \frac{C_2}{C_1}\sqrt{\left(\frac{L_2/L_1}{C_3/C_1}\right)}\right] \quad (65)$$

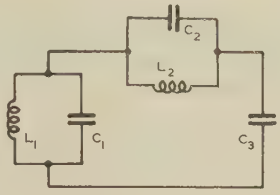


Fig. 17.—Equivalent circuit most commonly used in papers on third-harmonic tuning, obtained by neglecting the primary inductance kL_0.

and if $\omega_2/\omega_1 = 2{\cdot}80$,

$$\hat{V}_{C3} \simeq -1{\cdot}55\omega_1 V_0 \frac{T_s}{2}\left[1 - \frac{L_2}{L_2 + L_3} - \frac{C_2}{C_1}\sqrt{\left(\frac{L_2/L_1}{C_3/C_1}\right)}\right] \quad (66)$$

Eqn. (65) can be derived directly from eqns. (53) and (61). Its derivation as above highlights the separate effects of L_3 and C_2.

(7.6) Driver-Tube Voltage

As in the simple analysis of Section 5.4, the driver-tube voltage pulse is minimum for

$$\frac{B\omega_2}{A\omega_1} \simeq 0{\cdot}2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (67)$$

when the peak is given by

$$\hat{V}_{C1} = -0{\cdot}75\omega_1 V_0 \frac{T_s}{2} \quad . \quad . \quad . \quad . \quad (68)$$

(7.6.1) **Value at Mid-Point of Flyback.**

For the sake of completeness, it is recorded that the driver-tube voltage at $t = \frac{1}{2}T_f$ is given by

$$V_{C1} = -\tfrac{1}{2}V_0 T_s(A\omega_1 - B\omega_2)$$

Expansion of $A\omega_1 - B\omega_2$ gives

$$V_{C1} = -\frac{V_0 T_s \omega_1\omega_2}{2(\omega_2 - \omega_1)}\left[\frac{A\omega_2^2 + B\omega_1^2}{\omega_1\omega_2} - (A+B)\left(\frac{\omega_2}{\omega_1} - 1 + \frac{\omega_1}{\omega_2}\right)\right] \quad (69)$$

and substitution from eqn. (55) yields

$$V_{C1} = -\frac{V_0 T_s \omega_1\omega_2}{2(\omega_2 - \omega_1)}\left[\frac{N_0}{\sqrt{D_0}} - \left(\frac{\omega_2}{\omega_1} - 1 + \frac{\omega_1}{\omega_2}\right)\right] \quad (70)$$

If $\omega_2/\omega_1 = 2{\cdot}80$, eqn. (70) becomes

$$V_{C1} = -1{\cdot}55\omega_1 V_0 \frac{T_s}{2}\left(\frac{N_0}{\sqrt{D_0}} - 2{\cdot}16\right) \quad . \quad . \quad (71)$$

When the driver-tube pulse is flat topped, the maximum voltage is

$$\hat{V}_{C1} = -0{\cdot}75\omega_1 V_0 \frac{T_s}{2}$$

occurring at $t = \frac{1}{2}T_f$. Therefore, from eqn. (71), a transformer of optimum design has

$$\frac{N_0}{\sqrt{D_0}} = 2{\cdot}65 \quad . \quad . \quad . \quad . \quad . \quad (72)$$

(7.7) Losses

If losses are included in the equivalent circuit, it is not possible to have a symmetrical flyback. The whole mechanism of the

solution breaks down, and, in general, ringing cannot be made to cease entirely at the end of the flyback.

This is confirmed in practice, for line output transformers always show some ringing.

(8) ACKNOWLEDGMENTS

The paper is based on research carried out in the Electrical Engineering Department of the University of Melbourne, 1957–58. This work was made possible by a Commonwealth Research Grant, which the author gratefully acknowledges.

In addition, the author wishes to thank all those who in any way assisted him in his work, in particular Mr. A. E. Ferguson who read the manuscript of the paper.

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* The author's attention was drawn to this paper after he had submi revised manuscript. The experimental results give an excellent confirmation theoretical predictions.

SOME REMARKS ON WAVEGUIDE MODES

By R. A. WALDRON.

(*Communication received 2nd December*, 1960.)

Professor Barlow suggests[1] that, in obtaining a solution of the electromagnetic wave equation, it would be helpful to attempt first to resolve the field equations into three parts, corresponding respectively to TEM waves, surface waves, and waveguide modes. Such an attempt is not likely to be fruitful, because there is no real distinction between these waves, either mathematical or physical.

Before explaining this in more detail, it is necessary to introduce the concept of the homogeneous simple perfect (h.s.p.) waveguide.[2, 3] By 'waveguide' is meant a system in which all boundary surfaces are parallel to a given straight line. A homogeneous guide is one with only one boundary surface, on one side of which is a perfect conductor and on the other a single homogeneous medium. A simple guide is one in which all the media have only scalar properties. A perfect guide is one in which all the media are lossless. The h.s.p. waveguide possesses all these properties; it consists, therefore, of a perfect conductor enclosing a straight hollow space of arbitrary cross-section, filled with a lossless, scalar medium. In a special case, this might be vacuum. It will be convenient in the present note to take the h.s.p. guide to be of circular cross-section, although this is not necessary.

It can be shown that in general there is a one-to-one correspondence between the normal modes of any waveguide and those of the h.s.p. guide. Consider, for example, an ordinary practical guide consisting of a metal tube filled with air. This may be regarded as a guide with a single boundary surface of circular cross-section, inside which is a medium of relative permittivity $\epsilon_1' - j\epsilon_1''$ and outside which is a medium of relative permittivity $\epsilon_2' - j\epsilon_2''$, where ϵ_1' is nearly unity, $\epsilon_1'' \ll 1$, and ϵ_2'' is very large. This waveguide can be reduced to the h.s.p. guide by changing ϵ_1'' to 0, ϵ_1' to 1, ϵ_2' to 0, and ϵ_2'' to infinity. It is possible to solve Maxwell's equations for the metal-tube guide, and to obtain a set of eigenvalues (the phase constants) and eigenfunctions (the field components); as the guide is reduced, mathematically, to the h.s.p. guide, the eigenvalues and eigenfunctions also reduce to those appropriate to it. The cess can be carried out continuously, and at no point do any modes appear, nor do any disappear. There is thus a mo the metal-tube guide corresponding to each mode of the h guide, and vice versa. The metal-tube guide has been disc in more detail in Reference 3. An analogous reduction ca carried out for any non-h.s.p. guide; for example, if the gui cylindrical and contains a concentric dielectric rod, the redu consists in either reducing the radius of the rod to zer changing its dielectric constant to unity. The cases of a c drical guide containing a concentric rod of ferrite or a tu ferrite whose outer radius is equal to that of the guide have treated in References 4 and 5.

By treating the coaxial line as a waveguide it is possib use the above type of reduction to show that the transmis line mode (Barlow's TEM mode) is related to certain o modes of the h.s.p. guide. This can be done by general somewhat the mathematical problem; the system consider a guide with two boundary surfaces $r = a$, $r = b$, with a The outermost region, $r > a$, contains a perfect conductor region $a > r > b$ contains vacuum; and the inmost re $r < b$, contains a scalar dielectric with relative permittivi If we put $\epsilon = -j\infty$, the system becomes an ideal co transmission line; if $\epsilon = 1$, the system is an h.s.p. guide. T is a class of modes which, on reducing ϵ to unity, reduce t E_{PO} modes of the h.s.p. guide. In the textbooks we are that the guide of circular cross-section does not support modes because for these the boundary conditions can on satisfied if all the field components vanish identically. is true, however, only for the h.s.p. guide; as soon as ϵ de from 1, non-trivial solutions of Maxwell's equations be possible. At the other extreme, when ϵ goes to $-j\infty$, the modes degenerate to a single mode, the transmission-line m This question is discussed further in Reference 3.

Surface waves can be examined in the same way, and 'su wave' modes can be related to those of the h.s.p. guide. example, Clarricoats[6, 7] has related the 'surface-wave' mod

Mr. Waldron is with Marconi's Wireless Telegraph Co., Ltd.

ectric rod to those of a guide of circular cross-section ning a concentric dielectric rod. One further step, as bed above, would relate them to the modes of the h.s.p.

Alternatively, we could write ϵ_1 for the dielectric con- of the rod and ϵ_2 for that of the medium outside. By ing ϵ_2 continuously from 1 to $-j\infty$ and ϵ_1 to 1, the dielec- d guide would be reduced to the h.s.p. guide. This l bring out the absence of any meaningful distinction en 'surface-wave' modes and waveguide modes, for at no in the reduction could it be said that a change occurred essential nature of the waves.

correct approach to a new guided-wave problem is tly not to try to separate out the modes as transmission- surface-wave or waveguide modes, which might introduce ls, but to seek to relate the mode spectrum of the system investigation to that of the h.s.p. guide. In this way, ambiguous system of mode nomenclature can be arrived The mode in the new system which reduces to, say, the mode of the h.s.p. guide can be designated the H_{PQ} mode ut further ado. When this is done, the distinction between mission-line modes, surface-wave modes and waveguide s is seen to be quite artificial.

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[Professor H. E. M. Barlow makes the following comments: The important point made by Mr. Waldron, that there is in fact a continuous transition from one mode to another in accordance with the relative nature of the supporting media and the corresponding boundary conditions, emphasizes the difficulty of distinguishing clearly between different modes, but it does not in my opinion entirely preclude that possibility.

We are in fact accustomed to recognizing the TEM wave and the so-called waveguide modes as capable of excitation when specific conditions are established. In the first case there is no phase change over the wavefront, whereas in the second a standing wave transverse to the direction of propagation is essential. The surface wave is more difficult to identify because in some respects it is very similar to the TEM wave, but the important feature is the evanescent distribution of field over the wavefront. All of these waves in their pure form only transmit power in the direction of propagation, and recognizing their individual characteristics they should be capable of identification mathematically as well as physically. The fact that in practice we invariably have to deal with hybrid waves would seem to emphasize the value of separating components in terms of clearly established conceptions of the distinctive types of wave modes.

This is perhaps another way of saying much the same thing as Mr. Waldron, when he suggests that the right procedure is to relate the mode spectrum of the system of interest to that of the h.s.p. guide. To me the difference lies largely in the extent to which the concepts yield familiar interpretations in terms of wave modes of practical concern.]

DISCUSSION ON

'A CORRELATOR EMPLOYING HALL MULTIPLIERS APPLIED TO THE ANALYSIS OF VOCODER CONTROL SIGNALS'*

essrs. J. N. Holmes and J. N. Shearme (*communicated*): The ors describe a satisfactory correlator for computing the t-term correlation functions of low-frequency signals, but examples of the computed correlation functions which they v give rise to some doubts about the measurement procedure. e have studied the paper by R. E. Burgess (the authors' rence 8), which deals with envelope-detected band-limited e, both with and without the presence of an added carrier. n this paper the following results emerge:

(i) Both linear and square-law rectification give very similar sults for the power-density spectrum, and hence the auto-correla- on function, of the detected noise. For this reason it will be rmissible to consider only the square-law case.

(ii) The normalized auto-correlation function for detected noise one is

$$\rho(\tau) = \left(\frac{\sin \pi B\tau}{\pi B\tau}\right)^2$$

here B is the bandwidth of the noise in cycles per second.

Billings, A. R., and Lloyd, D. J.: *Proceedings I.E.E.*, Paper No. 3295 E, Sepber, 1960 (107 B, p. 435).

(iii) For the authors' Fig. 6 (noise and carrier) the auto-correlation function may be easily calculated from the spectrum. The result is

$$\rho(\tau) = \frac{2}{3}\left(\frac{\sin \pi B\tau}{\pi B\tau}\right) + \frac{1}{3}\left(\frac{\sin \pi B\tau}{\pi B\tau}\right)^2$$

These auto-correlation functions apply to the noise output of the envelope detector after the d.c. spectral line is removed. If this d.c. component is not removed the correlation functions will have a constant component in addition to those of the forms shown above. The authors do not state whether they removed the d.c. component before correlation, but as the curves of their Figs. 5 and 6 tend to zero with increasing τ, it is assumed that this component was not present.

Functions (ii) and (iii) above differ considerably from the authors' Figs. 5 and 6. Three examples will suffice to illustrate this. Figs. 5(*a*) and (*b*) should be of identical shape and should differ only in time scale. Fig. 5(*b*) should never be negative. The ordinate of Fig. 6 should be $-0{\cdot}128$ at $\tau = 44$ millisec.

These disagreements may be explained by differences between

the authors' signals and the ideal signals assumed by Burgess, and perhaps partly by errors in calibration of the correlator. Whatever the cause of the discrepancies, however, it does not seem reasonable to claim agreement with the results predicted by mathematical analysis.

Unless their meaning of the phrase 'occupies a bandwidth' is very different from that generally accepted, the authors' conclusion drawn from the comparison of Figs. 7 and 8 is incorrect. A reasonable conclusion would be that the vocoder control signals for the speech input have more energy at low frequencies relative to the high frequencies than they do for a noise input. There is no evidence that the significant energy components all lie in a bandwidth of less than 25 c/s. It is quite possible that low-level components in the region of 25 c/s, not evident from a casual inspection of the correlation function, may have a considerable effect on the transmission of the speech sounds used for the authors' analysis.

The cross-correlation functions shown in Fig. 9(*a*) and (*b*) have not been normalized by the usual process for cross-correlation functions. As they stand, the curves give no indication of the degree of correlation, which, from the available evidence, may even be very small everywhere. The usual form of normalized cross-correlation function is

$$\frac{\phi'_{ab}(\tau)}{[\phi'_{aa}(0) \cdot \phi'_{bb}(0)]^{1/2}}$$

Fig. 9(*c*) is an even more striking example of an unusual normalization process. The function

$$\frac{\phi'_{ab}(\tau) - \phi'_{ab}(0)}{\phi'_{ab}(0{\cdot}1) - \phi'_{ab}(0)}$$

is necessarily zero at $\tau = 0$ and unity at $\tau = 0{\cdot}1$ sec, irrespe of the form of $\phi'_{ab}(\tau)$. There is thus no justification for st that Fig. 9(*c*) 'shows a pronounced delayed correlation'.

Professor A. R. Billings and Dr. D. J. Lloyd (*in reply*): Me Holmes and Shearme are quite correct in assuming that absence of a constant term implies its removal: in justific for this it may be said that the correlation function for d current is singularly uninteresting. For this reason it removed subsequent to measurement and before normal the results. As explained in the paper, the reason for measu the correlation functions was to obtain an indirect measu the self- and cross-power spectra. The relative spectral d butions, excluding direct current, are then of principal int and in this context the method of normalization is irrele However, we reassert that there is a pronounced delayed co tion indicated in Fig. 9(*c*), and indeed this delay has been de for more than one speaker.

The criticism that the measured correlation functions a cardinal functions, or combination of cardinal functions, naïve one. It was never claimed that the shape of the pass-band was rectangular, and a moment's reflection indicate the impossibility of this. As to the spectral occu of the control signals, corresponding to Fig. 8, the pr of periodic or quasi-periodic components at frequencies of a 10 c/s would be indicated by rapid changes in slope observed correlation function: again there is no evidence We would suggest that an acceptable definition of banc is the r.m.s. width of the Fourier transform of Figs. 8(*a*) a which is of the order of 10 c/s. This is not to suggest t components occur above this frequency, but only that th small.

DISCUSSION ON

'ANALYSIS OF A FREQUENCY-MODULATED CONTINUOUS-WAVE RANGING SYSTEM'*

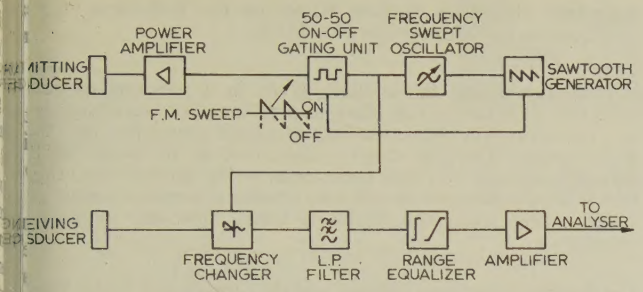
Mr. L. Kay (*communicated*): The paper can be misleading since the introduction suggests a generalized analysis wherein no approximations are made, but in fact the main context and conclusions are based upon a particular system developed for ionospheric sounding, with an approximation at the commencement of the analysis which reduces the value of the remaining theoretical presentation. Since Keep, Tucker and myself are quoted, each dealing with f.m. echo ranging systems (I think *echo* ranging is to be preferred since there are other methods of ranging not requiring a target echo) all of quite different types from that outlined by Messrs. Hyman and Lait, the link between the ionospheric sounder, radar echo-location, and sonar echo-location could perhaps have been made more clear. I cannot agree with the authors that all f.m. echo-location systems have been similar to airborne altimeters. This may generally apply, up to the present, in the case of radar, but certainly not in sonar. The authors should therefore have made it clear that their paper, whilst of general interest from the point of view of the method of mathematical analysis, applies mainly to a particular problem.

The approximations made at the commencement of the analysis assume that the frequency increases linearly with time for a period T_s followed by an instantaneous flyback. It is odd this approximation should be made when comparable app mations are not allowed in the remainder of the text. Ty values for T_s and α are not given, from which the validi the approximations may be judged, but taking a possible b width of 20 kc/s and a time interval of 1 millisec*, i.e. $T_s\Delta f$ = the linearity is far from that required to maintain the sp lines of the beat notes constant for integration over a time interval. This fact—a most important one—appea have been forgotten; the authors' views on this will be a ciated.

One of the main features of the paper is the use of the ch switching system to reduce the effects of range ambiguity, is certainly serious in the form of frequency-sweep descr This is not the only form of sweep which can be used, hov I have used a transmission consisting of a frequency s covering the maximum bandwidth of the array and lastin a period $T = 2R_m/c$, followed by a 'silent' period also of dur T, to advantage. During the silent period the f.m. osci was allowed to continue sweeping for a further change Δf to the array bandwidth. The oscillator sweep thus co twice the transmitted bandwidth.

* HYMANS, A. J., and LAIT, J., Paper No. 3264 E, July, 1960 (see 107 B, p. 365).

* Reference 1 in paper.



.—F.M. echo-location system with reduced range ambiguity effects.

ead of relying on the pick-up from the transmitter array n. oscillator was fed to a frequency changer in the receiver ig. A) so that a difference frequency was produced which d less from range ambiguities than even a pulse system. nly disadvantage is that all signals last for only 50% of the of observation with a consequent loss in signal/noise ratio B compared with an ideal fully continuous transmission n.

s possible (with added complication) to produce a fully uous beat note—rather than an intermittent one—from a single-line spectrum can be obtained, but an explanation h a system is more in keeping with a paper than a discus- I have not found such a complication justified in applica- with which I am concerned, but there are clearly applica- where it can be of value.

underwater sound applications the theoretical limitations ge resolution given in the paper are of little practical value, a resolution of several thousand units would be required to these limits. For example, operating at 100 kc/s covering dwidth of transmission of 10 kc/s and working to a maxi- range of 500 ft, the number of units of range resolution ole would be $T\Delta f = 2000$, which is far in excess of the al requirements of such systems.

vill also be observed that there would be little difficulty due ine spectrum rather than a continuous spectrum under such circumstances, since a filter would normally embrace several lines. When using a 50% 'on' and 50% 'off' transmission it is possible to shape the leading and trailing edges of the transmission pulse in such a way that spectral overlap due to two targets is very small; i.e. the higher harmonics of the repetition period are much smaller than for a rapid rise and fall of signal amplitude.

Messrs. A. J. Hymans and J. Lait (*in reply*): We thank Mr. Kay for his observations, but would point out that in our paper no claim to complete generality was in fact made. As stated in the introduction, the paper arose from design studies for an ionospheric sounder and the parameters involved may differ considerably from those used by Mr. Kay in sonar investigations. Nevertheless, the arguments have intentionally been kept general in nature, so that extension to other equipment having parameters of a different order may be possible. Our main claim is that certain assumptions made by Gnanalingam have been rendered more general.

As far as the maintenance of sweep linearity is concerned, the major problem is of a practical kind and was therefore omitted from our paper, which set out to discuss 'some aspects' from a theoretical point of view. It is obviously very important that linearity of sweep should be maintained, and we have given considerable thought to the question during design and construction. To put the problem in truer perspective, our T_s is 140 millisec and our bandwidth greater than 25 kc/s at all stages. The product of these parameters is thus many times greater than the figure of 20, which Mr. Kay rightly says would be inadequate.

The choice of system parameters is indeed not only important inasmuch as sweep linearity is concerned; noise problems, for instance, are equally relevant, and these are considerations which have been examined in a further paper.*

We thank Mr. Kay for drawing our attention to alternative methods for avoiding range ambiguity. The question of range resolution is again one of general interest which was treated here for completeness. It would obviously be inapplicable to a sonar system with the parameters quoted by Mr. Kay.

* Lait, J., and Hymans, A. J.: 'The Output Spectral Density of a Detector operating on a F.M. C.W. Radar Signal in the presence of Band-Limited White Noise', *Proceedings I.E.E.*, Monograph No. 412 E, October, 1960 (**108** C, p. 197).

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ress Report on the Development of a Photo-Electric Beam-Index Colour-Television Tube and System. Paper No. 3468 E.

R. Graham, M.Eng., J. W. H. Justice and J. K. Oxenham, M.A.

e paper is a progress report on the development of an index ur-television display using photo-electric indexing, and outlines of the basic problems associated with index systems. The effect oss-modulation between the writing and index signals is examined a method of overcoming this cross-modulation is suggested. The t of the delay around the index loop and its relationship with the horizontal time-base linearity is discussed, and a preferred circuit arrangement which relaxes the tolerances on this linearity is described. Some comparisons are drawn between index and other colour-television displays, and details are given of the practical verification of the system described. The present system may lead to a simplification of the colour display tube.

Topological Synthesis of Non-Reciprocal Resistance Networks. Monograph No. 428 E.

R. Onodera.

The paper deals with the topological synthesis of non-reciprocal resistance networks specified by the so-called driving conductance matrix. Synthesis of non-reciprocal networks has already been treated by H. J. Carlin and Y. Ooono, and their networks are specified by scattering matrices and consequently *n*-port impedance matrices. One of the merits of this paper, which differs from theirs, is in the determination of the connection and ideal transformers. Factors which determine the network are the connection and element resistances, and they therefore make possible the determination of the node-

branch incidence matrix and resistances so that the driving conductance matrix may equal the given matrix. When we denote the branch e.m.f. matrix by V and the branch-current matrix by I, $I = GV$, where, in general G' equals G, G is called the driving-conductance matrix, which is assumed to be given. If a maximal set of solutions for $Gx = O$ and $x'G = O$ is denoted by X as the matrix expression, X is called the zero factor of G. Provided that the zero factor X satisfies the character of a node-branch incidence matrix, the connection can be fixed directly. If not, the connection is determined by the part of X which satisfies the character of a node-branch incidence matrix, and the ideal transformers needed are determined by the other part. After the connection and the ideal transformers are determined, the branch resistances are determined by comparing the given matrix with the matrix calculated from the newly determined connection and the unknown branch resistances.

The Impulse Response of a Number of Identical Circuits in Cascade. Monograph No. 429 E.

K. THARMALINGAM, B.A., B.Sc.

Approximate analytical expressions are derived for the impulse response of identical circuits in cascade. The results obtained from these approximate expressions, tested for an exactly solvable case, show remarkable accuracy. Cruder approximations to obtain a quick estimate of peak-to-peak ring in response to a step function, and the first zero of impulse response, are also given.

Microwave Propagation through Round Waveguide Partially Filled with Ferrite. Monograph No. 430 E.

A. J. BADEN FULLER, M.A.

The theory of microwave propagation through an unbounded ferrite medium magnetized along the direction of microwave propagation is considered and expressions are derived for the components of the microwave field. Boundary conditions are applied, and characteristic equations are derived for four circularly symmetric shapes, namely

(*a*) Ferrite-filled round waveguide.
(*b*) Ferrite rod at the centre of round waveguide.
(*c*) Ferrite tube adjacent to the wall of round waveguide.
(*d*) Ferrite tube in the centre of round waveguide.

Some propagation coefficients have been computed for the last three shapes. The method is such that it may easily be extended to give the characteristic equations for any circularly symmetric shape of ferrite inside round waveguide in addition to the characteristic equations given in the paper. At 9370 Mc/s some computed results of Faraday rotation have shown reasonable agreement with those measured on ferrite rods. The results of a specimen investigation of the variation of the propagation coefficient with change of ferrite dimension are included.

Microwave Hall Effect and the Accompanying Rotation of the Plane of Polarization. Monograph No. 431 E.

Professor H. E. M. BARLOW, B.Sc.(Eng.), Ph.D.

Faraday rotation of the plane of polarization of an electromagnetic wave, propagated through a medium in the direction of an independently applied steady magnetic field, has been interpreted in terms of gyromagnetic action and in certain circumstances as a consequence of Hall effect.

The paper calls particular attention to the Hall-effect mechanism, which is shown to lead to different propagation coefficients for the right-handed and left-handed circularly-polarized components of a plane wave.

The contention is advanced that a Hall effect can be expected to arise from a displacement current in the material medium and that, as a consequence, the Faraday rotation effect in dielectrics and semiconductors may, in some cases, be significantly modified at frequencies above the microwave part of the spectrum. Some elementary calculations are given in support of this argument.

A Method of Calculating the Transfer Functions of Ladder Networks. Monograph No. 432 M.

N. REAM, B.A.

From the Kirchhoff equations applied to a ladder network, recurrence equations are derived for calculating the voltages and currents in terms of the output voltage and current; from these, another set of equations is obtained for working from the input end of the network. It is shown that either set of equations is particularly suitable for the numerical calculation of transfer functions. A few general formulae are included for the voltage-transfer ratio of an *RC* ladder network.

The Part Played by Surface Waves on the Reflection at a Fe Boundary. Monograph No. 433 E.

L. LEWIN.

It is shown that the solution given in a previous paper for reflection at a ferrite boundary in a rectangular waveguide is in in a certain region because it implies a finite power flux into the w guide walls. This flux can be taken care of by means of Bres surface waves, which can propagate in the anomalous region, which were postulated in order to resolve a thermodynamic para Their reality when the ferrite is lossy, however, requires fu elucidation.

The Surface Impedance Concept and the Structure of Radio Waves Real Earth. Monograph No. 434 E.

Z. GODZIŃSKI.

The paper is concerned with the concept of surface impedan applied to the theory of radio-wave propagation over a rea inhomogeneous and irregular, earth; some of the conclusions also be of value in the theory of waveguides, cavity resonato certain types of aerial.

The advantages and limitations of the surface-impedance co are shown in connection with a very general integral equation fo field strength.

The approximations and physical phenomena underlying the su impedance concept are first discussed in the simplest case of a h geneous and flat earth. The analysis is then extended to a horizon stratified earth; it is then possible to characterize the earth by co effective parameters depending on frequency and the geo structure of the soil. The same is also true in the case of a spi or not too excessively irregular earth. The discussion is generalized to the case of an arbitrarily inhomogeneous earth. height/gain function and the shape of the ellipse of polarizatio discussed. Approximate boundary conditions for the Hertzian ve and Hertzian scalar function are related to the concept of s impedance.

Practical conclusions are drawn with regard to the existence measurement of the effective earth constants and to some aspec geological prospecting by radio methods.

The Analysis of Non-Linear Resonant Circuits. Mono No. 435 M.

C. B. NEWPORT, Ph.D., and D. A. BELL, M.A., Ph.D.

The paper describes two graphical methods of calculating instantaneous voltages and currents which result from feedin sinusoidal current to a resonant circuit which employs a non-li inductor. The voltage-integral method uses a step-by-step approxi tion to the actual inductor characteristic, including hysteresis. Kn ledge of the detailed waveforms facilitates the determination of optimum magnitude and timing of a pulse to trigger a ferroreso circuit from the dissonant to the resonant state. Calculated wavefo are satisfactorily confirmed experimentally.

Alternatively, the phase-plane method allows the use of a piece linear approximation to the magnetization characteristic of the of the inductor (e.g. three straight lines), but at the cost of igno hysteresis.

Design of Cylindrical Surface Waveguides with Dielectric and Magn Coating. Monograph No. 436 E.

T. BERCELI, Dr.Ing.

The paper deals with wires whose coating has both dielectric magnetic properties. Approximate formulae are given for the concentration, propagation time, group velocity, power distribu characteristic impedance, losses and maximum transmissible powe such coated wires.

The magnetic coat concentrates the field more strongly than the dielectric one. The field concentration increases quickly increase of permeability and hardly depends on permittivity.

The attenuation is due to the conductive, dielectric and magn losses of the guide. The dielectric loss depends also on permeabi and vice versa. The total loss of a given transmission line as a func of field concentration has a minimum, which can be achieved coating of appropriate thickness.

To determine the maximum transmissible power, two formulae be used. One determines the maximum transmissible power the highest permissible value of electric field strength, and the o from the highest permissible degree of overheating.

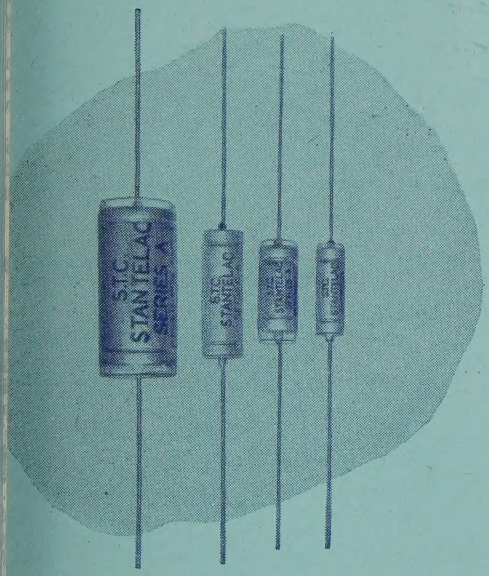
The paper concludes with a description of design procedure surface waveguides, illustrated by an example.

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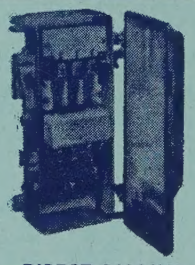
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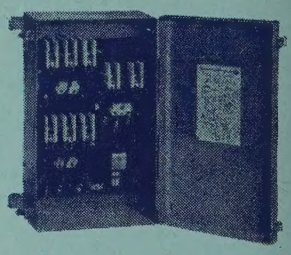
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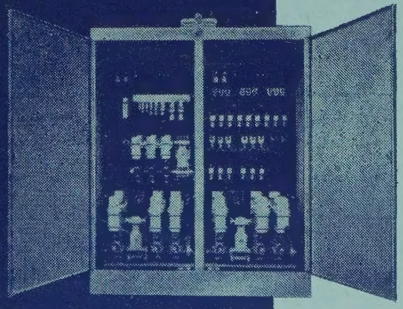


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